

UNIVERSITY OF ALBERTA

**Transmission Loss and Dispersion in Conductor-Backed Coplanar
Waveguide**

An Investigation of Waveguide Structures for Broadband Optical Modulators

by

Haidong Sun



A thesis submitted to the Faculty of Graduate Studies and Research in partial fulfillment
of the requirement for the Master of Science degree

Department of Electrical and Computer Engineering

Edmonton, Alberta

Spring 2003



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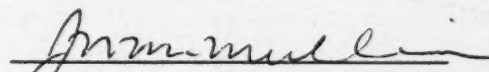


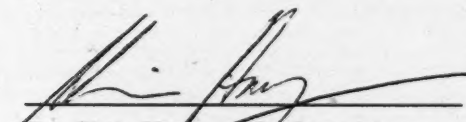
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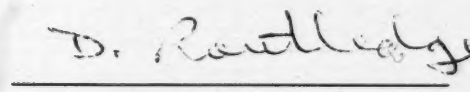
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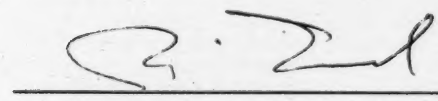
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For my family

ABSTRACT

The Conductor-Backed Coplanar Waveguide (CBCPW) as a planar transmission line structure is immune to dielectric loss in the substrate due to its conductor-backed configuration, therefore compatible with many substrate systems. This flexibility is very attractive for silicon-based applications where dielectric loss is a big problem due to its poor intrinsic resistivity. The CBCPW structure is proposed as the electrode structure in a low cost silicon-based electrooptical polymer modulator. However, it is possible that the parallel-plate structure formed by the upper and lower conductor layers of CBCPW gives rise to parallel-plate leaky losses.

This report is focused on the investigation of various loss mechanisms and dispersion effect in CBCPW by matched-through-line and resonance method. A transmission loss of 4.9dB at 20GHz is measured in a 50Ω CBCPW through line of 3.4617cm long micro-fabricated on silicon wafer, which suggests a modulator built with CBCPW electrodes could achieve half power modulation if velocity match between the electrical and optical signals is achievable at this high bit rates. A very small dispersion effect is observed in CBCPW across a wide frequency range from 50MHz to 20GHz by resonance method, and the attenuation constants obtained by this method agree well with the result from the matched-through-line method. It is proved that conductor loss is the main source of transmission loss in CBCPW, and there is no noticeable parallel-plate leaky loss in the investigated frequency range.

ACKNOWLEDGEMENT

I would like to first acknowledge the support and guidance of Dr. Jim McMullin, Dr. Ray DeCorby and Dr. Chris Haugen as my supervisors during the period of my study and research at the University of Alberta and *TRLabs* Edmonton. I would like to thank David D. Clegg, the Lab manager at *TRLabs*, who gave me a lot of help in conducting microwave transmission line measurement and experiment setup. I appreciate the help from Hue Nguyen, and Mike O'Toole with Big Bandwidth in the transmission line micro-fabrication process. I also want to thank Dr. David Routledge and Dr. Richard Marchand for their work on my thesis. Finally I would like to thank all the staff members and students at *TRLabs* Edmonton who have been very supportive and providing all the conveniences to me.

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LIST OF SYMBOLS

c	propagation speed of light
v	phase velocity of a electromagnetic signal in free space
v_{cpw}	phase velocity of a electromagnetic signal in CPW
ϵ_0	dielectric permittivity of free space
ϵ_r	dielectric permittivity of a material
ϵ_{eff}	effective dielectric constant of a transmission line
μ_0	permeability of free space
μ	permeability of conductor
C	capacitance of a transmission line
C_{air}	capacitance of a transmission line with dielectric material replaced by air
Z_0	characteristic impedance of a transmission line
β_0	phase constant in free space
α_0	attenuation constant of a signal
σ	conductivity of conductor
δ	skin depth in a lossy medium

LIST OF ABBRIVIATIONS

2D	two dimensional
3D	three dimensional
BCB	B-staged bisbenzocyclobutene
CAD	computer aided design
CPW	coplanar waveguide
CBCPW	conductor-backed coplanar waveguide
EO	electrooptic
HRS	high resistivity silicon
IC	integrated circuit
MEMS	micro electro mechanical system
MIC	microwave integrated circuit
MMIC	monolithic microwave integrated circuit
MZ	Mach-Zehnder
Q_0	unloaded quality factor
Q_L	loaded quality factor
RFIC	radio frequency integrated circuit
RPM	round per minute
S-parameter	scattering parameter
TEM	transverse electromagnetic
T_g	glass transition temperature
UV	ultraviolet
VNA	vector network analyzer

1. Introduction

Since the 1970's we have seen a rapid development in fiber-optic communications due to its inherent advantage over copper transmission lines such as low cost, low loss, long distance transmission, huge bandwidth and immunity to electrical noise. This trend has been enhanced by the tremendous reduction in loss and non-linear effect in fibers. The record of lowest fiber loss so far is 0.151 dB/Km at a wavelength of 1568 nm achieved by *Z-PLUS Fiber*™ by SEI (Sumitomo Electric Industries, Ltd. Japan), which further extends the transmission distance of a single fiber before any amplification or signal processing is needed to reshape the transmitted signal. Fiber-optical communication components are under intensive investigation in hopes of further reduction in manufacture cost and improvement in device performance.

Today, many people in the industry are talking about "fiber to the home", which could provide small business and individuals with faster than ever access to information and entertainment resources on the World Wide Web. This is obviously a very attractive prospect to many who are tired of waiting on the jammed information highways. The number of expected users of Fiber-to-the-home (FTTH) systems in the United States will reach 2.65 million by 2006 [1]. But the vast bandwidth and high speed internet access provided by fiber can not be fully accessible to individual end users partly due to the high price of key optoelectronic components in the systems, such as optical sources, switches, and modulators, to name a few.

The situation is analogous to that of the computer industry. The huge computational power provided by computers could be offered to individuals only after

the cheap and massive production of silicon chips was realized. To reach such an ambitious marketing goal while providing affordable services to ordinary consumers, equipment manufacturers have to produce competitively priced products. Integrated optoelectronic components based on new material and technology are the focus of current research and development due to their inherent merits of high reliability and cost efficiency.

The objective of this study is to investigate the possibility of building an integrated electrooptic polymer modulator on silicon by using polymer materials and a Conductor-Backed coplanar Waveguide (CBCPW) structure with an emphasis on the study of transmission character of CBCPW structure.

Traditionally, high-speed optical modulators are built with semiconductor crystals such as Gallium Arsenide (GaAs) and Lithium Niobate (LiNbO_3) due to their electro-optic (EO) effect, which is an electronically induced change in refractive index of the material along its crystal axis by an external electric field. These materials are expensive to build and the fabrication process has to be stringently controlled for crystal growth. More over, the biggest problem is that they cannot be integrated with other available low cost silicon based components, and they have to be used as a stand-alone module, which degrades the system integration and reliability. With advances in material science and technology, electro-optic polymers have been considered as promising substitutions for these expensive crystals. These polymer materials possess a number of potential advantages in high-speed modulator applications [2]. The electro-optic coefficient of electro-optic polymers has reached the same level as that of LiNbO_3 [3], and the

flexibility in device fabrication makes polymer based devices compatible with silicon technology [4]. The most attractive merit for high-speed devices based on polymers is that the dispersion in polymers is very small compared to LiNbO_3 , which means a small velocity mismatch between the microwave and optical signals could be achieved and this leads to increased modulation bandwidth. EO modulators based on polymers with bandwidth of 20 GHz [5] have been demonstrated.

1.1. Principle of EO modulator operation

The top view of a symbolic structure of Mach-Zehnder modulator is shown in Fig.

1.1.

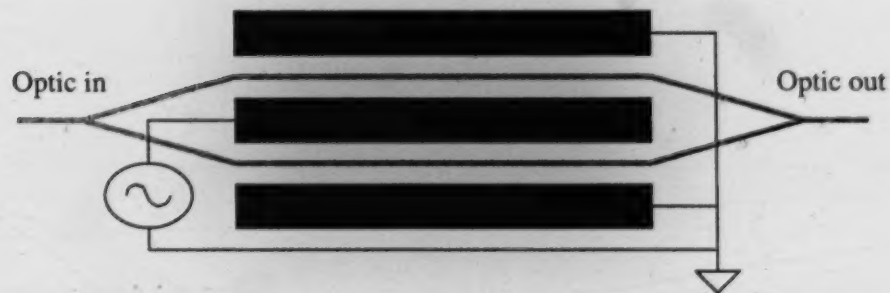


Fig. 1.1: Schematic of a Mach-Zehnder modulator in top view

It is built on a LiNbO_3 substrate and is composed of optical waveguide with EO effect and electric waveguide. One coherent optical carrier signal is equally split and passed through two optical arms that are sandwiched in the gaps of the electrodes and subject to a microwave-modulating signal on the electrode. Due to the EO effect induced by the electrical signal, the refractive index of the optical waveguide material can be increased or reduced depending on the direction of the applying electric fields relative to the optical axis of the optical waveguide. In the configuration above, if a positive voltage is applied

on the center conductor and the side conductors are grounded, the electrical fields across the two branches of the optical waveguide are in the opposite direction and so are the EO effects in the two optical arms. Thus the optical signals propagating in the two optical waveguides see the difference of the refractive indices and propagate at different speeds. The phase difference between the two optical signals could be exactly 180 degrees at the ends of the two arms where they meet if the length of the optical arms is properly designed, and a destructive interference of the two optical signals could be achieved at the output port. By alternating the electric signal, there is an alternation between constructive and destructive interference and a magnitude modulation of the optical signal is achieved.

The bandwidth-voltage ratio is often used in evaluating the performance of an EO modulator. The bigger the value the better the performance, which means maximum phase shift could be achieved by minimum electric potential. There are two major influencing factors that limit the achievable bandwidth of the modulator, namely velocity matching between optical and electric signals and electrical loss in the electrode. The operation of the modulator depends on the EO effect of the optical waveguide induced by electrical signal, it is desirable to have the optical signals experiencing the strongest electric field where they travel and this requires a velocity matching between electrical and optical signals. The speed of the optical signal in the optical waveguide is determined by the refractive index of the waveguide material at the optical carrier wavelength, while the velocity of the microwave modulation signal is determined by the effective dielectric constant of the transmission line electrode at the desired microwave frequency. When the effective dielectric constant of the transmission line at microwave

frequencies equals the square of the refractive index of the optical waveguide at the optical frequency, the electromagnetic field of the microwave signal can travel at the same speed as the optical signal and the maximum overlapping between the two signals can be achieved. This leads to the maximum modulation and phase shift of the optic signal in the optic waveguide. The electrical loss in the modulator electrode is caused by two sources, the material loss (both conductors and dielectrics) and radiation loss. The material loss, especially the frequency dependent conductor loss, is a major concern, since it decreases the modulation bandwidth by reducing the phase modulation of optical signal.

1.2. Project goal

The driving force behind the project is to use less expensive EO polymer material to build an EO Mach-Zehnder modulator on silicon, which increases the potential of integrating the modulator with other silicon-based device. Silicon technology is the most economic semiconductor manufacturing technology so far, and widely used as a cost effective platform for electronics applications. However modulators built on silicon could experience excessive dielectric loss due to the poor intrinsic resistivity of the material. The conventional CPW electrode structure in GaAs based device has to be changed to Conductor-Backed Coplanar Waveguide (CBCPW) as shown in Fig. 4.2.1.2 (c) to combat the microwave energy loss problem in the silicon substrate. However the conductor-backed configuration might give rise to the so-called parallel-plate leaky mode propagation and excessive transmission loss.

This report investigates the feasibility of employing Conductor-Backed Coplanar Waveguide (CBCPW) transmission line as the electrode structure for microwave signal transmission in silicon based EO polymer Mach-Zehnder modulator. Focuses are put on the investigations of microwave loss mechanisms and microwave dispersion characteristic of CBCPW transmission lines with BCB organic polymer as the dielectric layer. Design, simulation, fabrication and measurement analysis of CPW and CBCPW structures will be given in details. Hopefully this work will provide useful information and contribute to the prototype design of EO polymer modulator in the future.

1.3. Outline of thesis

This report starts with introduction of background knowledge and general information related to waveguide and transmission line in section 2, and proceed to specific quasi-static analysis of CPW and CBCPW structures and their transmission loss analysis in section 3. In section 4, transmission line fabrication, mask layout and transmission loss measurements using the matched-through-line method and the resonance method are described. The measured transmission loss data and effective dielectric constant are given and analyzed in section 5. Conclusions and remaining work to be done are summarized in section 6.

2. Background and theory

In modern telecommunication systems higher frequencies are deployed for signal transmission in order to obtain longer transmission distance and wider bandwidth. At GHz signal frequencies, where the wave lengths are comparable to the dimensions of the electronic components, the parasitic effects become large to an extent that the lead inductance and transit time problem could make the circuits useless [6]. So microwave frequency components that are basically miniaturized electromagnetic wave guiding and control structures are needed in place of wiring, transistors and ICs of low frequency electronics. In the design of these high-speed microwave-guiding structures, transmission efficiency and dispersion effect are the most important aspects to be considered. This section is devoted to general information on microwave guiding structures and related design considerations such as dielectric material selection and transmission loss.

2.1. Microwave guiding structures and CBCPW transmission line

Traditionally, there are 3 categories of microwave-guiding structures, namely, rectangular waveguides, coaxial cables and planar transmission lines. Rectangular waveguides can be used to handle relatively higher microwave power than other transmission structures, but they are bulky and expensive. Coaxial cable is broadband and convenient for many microwave applications, but difficult to fabricate into complex microwave network. Planar transmission lines have the advantages of simple structure, easy fabrication that only involves planar photolithography and metal deposition process,

and compatibility with silicon and small sized semiconductor devices in Microwave Integrated Circuits (MIC), so they are widely used in modern communication circuits.

There are many types of planar transmission line structures available for high frequency circuit applications, such as micro-strip line, strip line, slot line and coplanar waveguide. The employment of CBCPW as a microwave-guiding structure in this project is based on several facts related to the operational character of the modulator, such as the device configuration, loss and fabrication issues.

First of all, the operation of MZ modulator needs a type of transmission line that provide an electrical modulation structure such that opposite EO effects can be induced by opposite high frequency electric fields in the two parallel optical waveguides. In the CPW and CBCPW structures, the dominant bonded electric field distribution pattern is called CPW mode as shown below:

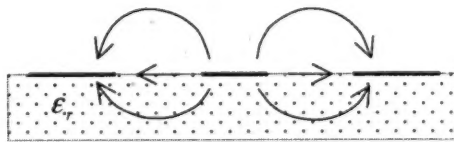


Fig. 2.1.1: Electric field distribution pattern of CPW mode

It provides desired opposite electric fields in the two gaps between the center and two ground conductors by applying an electrical signal on the center conductor and keeping the two side conductors grounded. The CBCPW structure enables push-pull operation of the device, which reduces the amplitude of required modulation signal by half.

Furthermore, the CPW structure can be used as an electrical polling tool for active polymer optical waveguides [7], which is an electric dipole alignment process in the polymer to produce desired EO effect in these optical waveguides. Second, since the

fabrication process of CPW involves printing and etching metal layer on only one side of the dielectric substrate, no vias and holes are needed for grounding purposes, and high inductive paths can be avoided.

Third, CBCPW structures provide higher mechanical stability and better heat sink ability that is desirable for the thermal stability of the dipole alignment in the active polymer for long-term device operation. The backside conductor plane also plays a significant role in preventing microwave power from leaking into lossy supporting silicon substrate. However the parallel-plate configuration formed by the backside conductor plane and top CPW conductor traces could give rise to undesired parallel-plate propagation mode as shown below,



Fig. 2.1.2: Electric field pattern of parallel-plate mode in CBCPW structure

It does not provide desired electric field distribution and represents power leakage and radiation from the dominant CPW propagation mode making the conductor-backed structure potentially dangerous to use [8]. This power leakage effect introduced by the conductor backing is a key investigating point in CBCPW structure.

2.2. Dielectric material and substrate (Silicon & Polymer)

Dielectrics can be defined as materials with high electrical resistivities that conduct virtually no electricity at low or dc electric fields. A large group of materials, including gases, liquids, semiconductors, ceramics, and organic and inorganic polymers,

are classified as dielectrics. However, there are no perfect dielectric materials, which means dielectric loss rooted from electron and ion movement cannot be avoided.

Dielectric substrates provide platforms to build electronic circuits and insulation between circuit blocks. They not only provide mechanical support to the circuits but also influence their properties, such as characteristic impedance, effective dielectric constant, dispersion and transmission loss. So one has to consider some important qualities of the dielectric substrate for microwave applications, which include:

- The microwave dielectric constant
- The frequency dependence of dielectric constant (material dispersion)
- The dielectric loss tangent, or imaginary part of the dielectric constant
- The surface finish and flatness
- The thermal expansion and conductivity
- The dimensional stability with time
- The surface adhesion properties for the conductor coatings
- The manufacturability (ease of cutting, shaping, and drilling)

The dielectric constant is the ratio of the capacitance induced by two metallic plates with an insulator between them to the capacitance of the same plates with air or a vacuum between them. The dielectric constant is used to determine the ability of an insulator to store electrical energy; the bigger the constant the better the electric field can be confined within the material. Optical fiber is a good example of confining the electromagnetic fields of a propagating signal in its core area by using high dielectric constant material as core and low dielectric constant material as cladding at near infrared

frequencies. The dissipation factor is defined as the reciprocal of the ratio of the insulating materials' capacitive reactance to its resistance at a specific frequency. It measures the efficiency of an insulating material, which is an analogy to the Q factor of a capacitor. The smaller the dissipation factor the less lossy the material is, and this quantity is a frequency dependent variable representing dielectric loss.

Silicon, as one of the most widely used semiconductor materials, possesses some of these desirable qualities. First, silicon technology is the most well developed semiconductor technology with the lowest cost. It provides a good platform for low cost electronic applications. Second, Silicon RFIC (Radio Frequency Integrated Circuits) are compatible with BJT and CMOS data processing and memory circuits, they could be monolithically integrated on a single chip. Further more, since sensors and MEMS (Micro-Electro-Mechanical System) devices are usually built on silicon substrates, single chip systems are possible, which means the packaging and interconnection costs and insertion loss can be significantly reduced while the overall reliability can be improved [9]. So silicon is considered with priority as a mature and cost efficient device platform.

The low cost and possibility of monolithic integration of devices on silicon are very attractive. However, one big problem with silicon devices is the electrical loss. Silicon has relatively big dielectric loss compared to other semiconductors due to its inherent poor resistivity, which is 10000 times less than that of GaAs. The signal attenuation related to dielectric loss in transmission lines built on ordinary low resistivity silicon is excessively high over the entire microwave and millimeter wave frequency spectrum [9]. It is a barrier for wider use of silicon in low loss RF applications. The

dielectric loss in silicon could be reduced by increasing the resistivity of the material, which gives rise to the name 'High Resistivity Silicon' (HRS) ($> 2500 \Omega\text{-cm}$). However, it has been shown that the loss term related to finite resistivity of silicon dominates dielectric loss through 50 GHz in CPW lines on silicon with the highest available resistivity of $10000 \Omega\text{-cm}$ [9]. Although CPW lines built on HRS can provide comparable attenuation to that of CPW on GaAs, the cost of such silicon substrates can be excessively high, because the material price increases with the increase in substrate resistivity. Several structural solutions have been proposed to combat the dielectric loss problem. The first one is to increase the distance between metal conductor and silicon substrate by adding other low loss material in the middle to keep the electromagnetic fields away from silicon substrate. The second is to add a metal layer on top of silicon to isolate microwave energy in the circuit from the lossy substrate. These two methods and some of the problems associated with them will be addressed later.

Polymer organic materials have long been predicted as promising materials in optoelectronic applications making dielectric buffer layer and optical waveguide. The most attractive merits of these plastic-like materials are low cost, compatibility with standard IC technology and flexibility in manufacturing. These promised merits of the material are major attractions to research workers who are looking for materials to build low cost optoelectronic devices, and a lot effort has been devoted to developing new polymer materials. The critical parameters of organic polymers for high-speed optoelectronic applications are:

- Optical transparency
- Refractive index

- Thermal stability under normal device operation
- Planarity and surface adhesion properties for the conductor coatings
- Glass transition temperature (T_g)
- Microwave property (dielectric constant)

High optical transparency is desired for polymer optical waveguides, to ensure that the optical signals are passed through the material with minimum loss. Thermal stability is a concern in active polymer material applications, where the polymer is doped by dipolar particles and the alignment of these dipoles could be relaxed by thermal motion of atoms in the material. Maximum surface planarity and adhesion to metal layer are desired which in turn reduces the surface roughness and electrical loss in the conductor layer. The glass transition temperature marks the temperature at which the material softens and becomes less supportive. It is desirable to find a material with high T_g ($> 200^\circ\text{C}$), which could stand relatively high temperature in device manufacturing processes. Microwave properties is another important consideration in design of high-speed electronics, where the dielectric constant is desired to be small to reduce the signal transmission delay.

The polymer material used in our project is called BCB-3022-63, which is a member of the dry-etch CYCLOTENE advanced electronics resins family developed for microelectronics applications by DOW Chemical [10]. The polymers are derived from B-staged bisbenzocyclobutene (BCB) monomers, and are formulated as high-solids, low-viscosity solutions. It is ideal for applications where a thin dielectric layer is required at the wafer level. Thin film coatings of $9.5\text{ }\mu\text{m}$ to $26\text{ }\mu\text{m}$ are achievable in a single spin-

coat application using standard IC processing techniques. Some of the noticeable properties of the polymer include:

- Low dielectric constant of 2.65 (Essentially independent of frequency)
- Low loss at high frequency
- Low moisture absorption
- High degree of planarization
- High optical clarity
- Good thermal stability and high T_g ($>350^\circ C$)
- Excellent chemical resistance
- Good compatibility with various metallization systems

The most attractive properties are: low dielectric constant, which usually means low dielectric loss and bigger dimensional tolerance in transmission line fabrication; Chemical resistance and high glass transition temperature making the dielectric more resistive to damages in the fabrication process after polymer spin coating process; A high degree of planarization ideal for supporting planar metal structures on top of it. Another advantage of choosing BCB polymer is that slab optical waveguide can be easily realized by polymers with slightly different dielectric constant in the same family. BCB polymers have been used in high-speed electronics packaging and interconnection applications [11-13].

2.3. Loss and dispersion effect in transmission lines

Any transmission line is subject to transmission loss due to the finite conductivity of the metal conductor, the dissipation and polarization of supporting dielectric material

and the radiation loss that incurs when the electromagnetic field of the propagating energy cannot be confined in the transmission line. These frequency dependent effects and the change in electromagnetic field distributions with wavelength together give rise to a dispersion effect, which describes the frequency dependency of the effective dielectric constant of the transmission line. To appreciate the effect, we can write the phase constant of a propagating wave as

$$\beta = \omega \sqrt{\epsilon_0 \epsilon_{eff} \mu_0} = \frac{\omega}{c_0} \sqrt{\epsilon_{eff}}$$

It represents the phase shift per unit length of a signal along the transmission line. If the permittivity of the transmission media varies with frequency, the phase constant is not a linear function of frequency. Different spectrum components of a broadband signal see different dielectric permittivities and propagate at different speeds, which leads to the dispersion or pulse broadening effect.

2.4. Considerations in transmission line design for the EO modulator

Transmission lines play an important role in analog MIC design. They are designed to transmit microwave energy with minimum loss. Since any discontinuity in the circuit would give rise to reflection and radiation loss, impedance matching at every connection nodes are needed and transmission lines act as impedance transformers where they are needed. A proper design of transmission line is a key to ensure the optimal operation of the whole circuit, transmission loss and dispersion have to be controlled in a tolerable range. The ease of implementation and fabrication are always blessings. Furthermore the selection of dielectric material as circuit substrate is also a key influencing factor in terms of circuit cost and performance.

3. Theory of CPW and CBCPW transmission lines

Characteristic impedance, effective dielectric constant, transmission loss and dispersion effect are the most important properties of a transmission line. They have to be carefully studied and modeled in order to evaluate the performance of these transmission line structures. In this section, these critical parameters in CPW and CBCPW are analyzed by quasi-static method and closed form solutions are given. The loss mechanisms in these lines are also analyzed.

3.1. Methods of transmission line analysis

Many numerical methods exist for planar transmission line analysis, generally they are divided into two main categories: quasi-static analysis and full wave analysis.

3.1.1 Quasi-static Analysis and conformal mapping technique

The fundamental mode of propagation in a two-conductor transmission line such as coaxial line is essentially a Transverse Electromagnetic Mode (TEM) when the conductors are considered to have infinite conductivity and the dielectric medium is homogeneous, lossless, and isotropic [14]. In practice such requirements are seldom satisfied; they usually have inhomogeneous dielectric or multi-layered structures, which cannot support pure TEM mode propagation [15]. However at low microwave frequencies (below X- band) where the strip width and the substrate thickness are much smaller than the guided wave length in the transmission line, the propagation mode closely resembles the TEM mode, and the longitudinal component of the electromagnetic field is small compared to those in the transverse directions, the non-TEM nature can be ignored. Then the 3-dimensional electromagnetic problem can be reduced to a solution

of 2-dimensional Laplace's equation subject to the boundary conditions that leads to a quasi-static transmission line analysis based on TEM mode assumption. In this method, conformal mapping techniques are used to transfer the physical structure of transmission line into a parallel plate structure, and the transmission line characteristic are calculated from two electrostatic capacitances of the structure, one capacitance C_{air} of a unit length of the transmission line configuration with the dielectric substrate replaced by air, and the other C for the line with the dielectric substrate presented. Values of the characteristics impedance Z_0 and the propagation constant β are given as:

$$Z_0 = \frac{1}{v\sqrt{CC_{air}}} \quad \beta = \beta_0 \sqrt{\frac{C}{C_{air}}} \quad \epsilon_{eff} = \frac{C_{air}}{C}$$

With v as the phase velocity of the electromagnetic waves in free space, and β_0 as the propagation constant in free space. Schwarz-Christoffel transformation is one of the most useful conformal transformation techniques for planar transmission lines, which involves using complex analytical functions to convert or map one mathematical problem from one domain into another under some very restrictive conditions. Since the deviation from TEM assumption is quite small in CPW line in the lower end of microwave frequency, the method can provide simulation accuracy comparable with that of full wave technique for frequency up to 40 GHz for most of CPW structures [16][17].

3.1.2 Other analysis methods and models

- **Dispersion models**

In this model, the deviation from the TEM nature of the propagation mode is accounted for quasi-empirically. The non-TEM nature causes the characteristic impedance and effective dielectric constant to be functions of frequency. The variation

of the characteristic of the transmission line with frequency is expressed by empirical solutions obtained by curve fitting and other approximation methods. Some parameters of the model are determined such that the final expression agrees with the known experimental dispersion behavior of the transmission line. Several dispersion models were given by different workers [18-20]; however there has not seen much work done for CPW and CBCPW lines. Further, since it is a semi-empirical method, the accuracy is highly dependent on the experimental set-up and subject to measurement errors. That is why this method is not popular for practical applications.

- **Full wave analysis**

Full wave analyses take into account the hybrid nature of propagation modes, which involves the longitudinal electromagnetic components, and cannot be fully described by the static capacitance and inductance derived from quasi-static analysis. One has to introduce time-varying electromagnetic fields and solve wave equations. The charge density used in the quasi-static analysis is replaced by electric current density to account the time varying nature of the method. In full wave methods the wave propagation constant is calculated instead of capacitance in the Quasi-TEM method and they are the most rigorous methods for transmission line analysis. The full wave analysis can be subdivided into frequency domain, space domain and time domain methods.

Compared with the complexity and computational burden imposed by full wave analysis, Quasi-TEM analysis is a simple and fast transmission line analysis method providing closed form solutions with good accuracy in the lower end of microwave frequency.

3.2. CPW and CBCPW model analysis using quasi-static method

For the Quasi-TEM approximation to be used in transmission line analysis, it requires the wavelength of the propagating signal to be large compared to the lateral dimension of the transmission line, which sets an upper limit for the working frequency of the device. In CPW line the distance between the two side ground conductors has to be kept much smaller than the guided wavelength [21].

A quasi-static analysis for CPW can be carried out by using a conformal transformation proposed by C. P. Wen [22], where the CPW line took an ideal form which consists of a conducting strip with two semi-infinite side conductors on the surface of an infinitely thick dielectric substrate and the conductor thickness was assumed to be zero. A zeroth-order quasi-static approximation was employed with the assumption that the slot between the center conductor and the side ground planes are magnetic walls where there is no tangential magnetic field and the dielectric half plane was transformed to the interior of a rectangle in an image plane by conformal mapping, and the CPW line characteristics were obtained by evaluating the line capacitance from the image plane in between a pair of transformed parallel metal plates.

After Wen's pioneer analytical analysis of CPW structures, the transmission line has gradually evolved from its original form in response to various technology development and engineering application requirements. There has been a lot of work [17, 23-29] on the study of CPW structures with various changes made to the original model, the most significant changes have been made to count the effects of finite thickness of the dielectric substrate, finite width of the ground electrodes and adding the

backside conductor plane. These changes are based on several facts, for example in modern MIC designs, semiconductors such as GaAs are used as 100 to 200 μm thick layer, which is far from the infinite thickness assumed in the original model. The electromagnetic fields may penetrate this thickness and propagate into other media if there is any underneath the semiconductor. Furthermore designers often add a metal layer on the backside of the thin and fragile semiconductor layer to increase the mechanical strength and improve power handling ability, which would cause a dramatic change in the field configuration in the vicinity of the back metal plane, and the parallel plate structure formed by the ground planes on either side of the center conductor and the backside conductor layer would cause energy radiations from the CPW mode if the electric potentials on the planes can not be kept at zero. These changes in CPW structures definitely need to be addressed when it comes to establishing design and simulation models.

One of the most important works on CPW analysis after Wen was carried out by Veyres and Hanna [30]. They analyzed CPW with finite dielectric substrate thickness and back conductor plane (Conductor-Backed Coplanar Waveguide) with conformal mapping method, and presented closed form equations for the model for the first time, which laid a foundation for multi-layer CPW analysis.

3.2.1 Single-layered CPW model without conductor backing

First, let us take a look at the original CPW model, and work through the analysis to get a set of closed form expressions for line impedance and effective dielectric constant by using quasi-static conformal mapping technique.

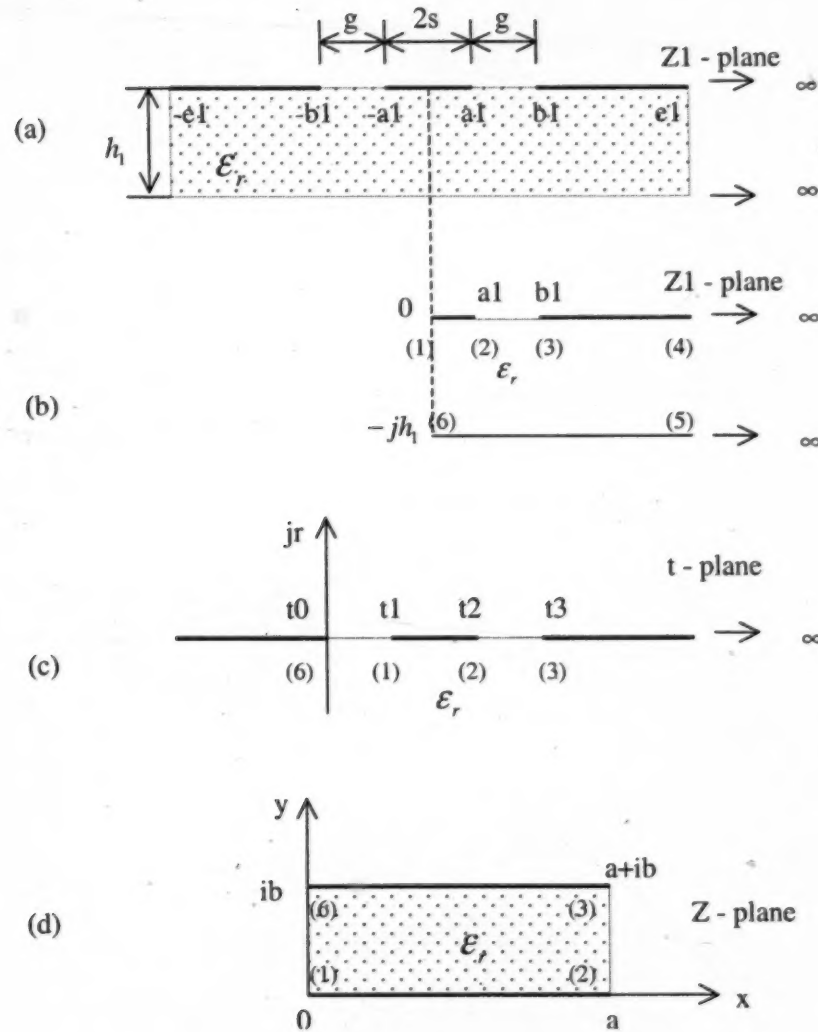


Fig. 3.2.1: Conformal Transformation of CPW line

The CPW to be analyzed is a symmetric structure shown in Fig. 3.2.1a. The width of the center conductor is $2s$, the gap between the center conductor and the ground planes is g , the two ground planes on either side of the center conductor are assumed to be extended to infinity and they are built on a dielectric substrate of an infinite thickness with dielectric constant ϵ_r . Since the magnetic field distribution around the center

conductor are largely perpendicular to the centerline of the structure, a vertical "magnetic wall" is assumed at $Z_1=0$. The two gaps are modeled as magnetic walls as well for the same reason, though this is rigorously true only when the line is immersed in a homogeneous dielectric. It has been proven to yield excellent results for practical line dimensions with reasonable gap width. [23]

We can consider the right half of the CPW shown in Fig. 3.2.1(b) out of the symmetry of the structure. By assuming the gap between the half center conductor and the ground plane as magnetic wall, the first conformal mapping transforms the right half of the dielectric substrate to the lower half of the t -plane using the Schwartz-Christoffel transformation [31] shown in Fig. 3.2.1(c).

$$t = \cosh^2 \left(\frac{\pi \cdot Z_1}{2h_1} \right), \text{ where } Z_1 = a_1 + jb_1. \quad (3.2.1-1)$$

So, we have

$$\begin{aligned} t_0 &= \cosh^2 \left(\frac{\pi \cdot (-jh_1)}{2h_1} \right) = 0, & t_1 &= \cosh^2(0) = 1, \\ t_2 &= \cosh^2 \left(\frac{\pi \cdot a_1}{2h_1} \right), & t_3 &= \cosh^2 \left(\frac{\pi \cdot b_1}{2h_1} \right). \end{aligned}$$

The second conformal mapping process is to transform the fields in the lower half of the t -plane into a finite image internal z domain using the following Christoffel-Schwartz transformation as shown in Fig. 3.2.1(d).

$$z = x + jy = a + jb = \int_1^t \frac{A \cdot dt}{\sqrt{t(t-t_3)(t-t_2)(t-t_1)}}, \quad (3.2.1-2)$$

where A is a constant and $Z=x+jy$ defines the transformed point from the Z_1 -plane. The integral solutions for the rectangle's coordinates in the Z plane are: [15]

$$a = A \cdot K(k_1) \quad b = A \cdot K(k_1') \quad k_1' = \sqrt{1 - k_1^2}, \quad (3.2.1-3)$$

where $K(k_1)$ is the complete elliptical integral of the first kind with modulus

$$k_1 = \frac{\sinh(\frac{\pi \cdot a_1}{2h_1})}{\sinh(\frac{\pi \cdot b_1}{2h_1})} \approx \frac{a_1}{b_1} = \frac{s}{s+g}, \text{ when } h_1 \rightarrow \infty. \quad (3.2.1-4)$$

Where a_1 and b_1 are dimensions in Z_1 -plane, and h_1 is the substrate thickness which is assumed to be infinite.

In Z -plane, the capacitance per unit length of the line contributed by the dielectric in the lower Z_1 plane can be written as:

$$C_1 = \epsilon_0 \cdot \epsilon_r \cdot (2a/b) = \epsilon_0 \cdot \epsilon_r \cdot 2 \frac{K(k_1)}{K(k_1')}. \quad (3.2.1-5)$$

The capacitance C_0 due to the free space half plane above the CPW structure can be obtained by mapping it into a finite rectangular area in Z plane in the same way as we did with the dielectric half plane following (3.2.1-1) – (3.2.1-4), but using h_0 in (3.2.1-1), and the capacitance C_0 can be written as

$$C_0 = \epsilon_0 \cdot 2 \frac{K(k_0)}{K(k_0')}, \quad (3.2.1-6)$$

where the modulus of the complete elliptic integral $K(k_0)$ is

$$k_0 = \frac{\sinh(\frac{\pi \cdot a_1}{2h_0})}{\sinh(\frac{\pi \cdot b_1}{2h_0})} \approx \frac{a_1}{b_1} = \frac{s}{s+g}, \text{ when } h_0 \rightarrow \infty \quad k_0' = \sqrt{1 - k_0^2}. \quad (3.2.1-7)$$

The height of the half free space h_0 can also be assumed as infinite compared to other physical dimensions in the structure.

The constant 2 appears in equations (3.2.1-4) and (3.2.1-5), because in the derivation only half of the lateral physical structure was transformed into Z-plane taking advantage of the symmetry of the structure, and the capacitance should be twice of that in Fig. 3.2.1(d). The total line capacitance per unit length including the free space half-space is

$$C = C_1 + C_0 = \epsilon_0 \epsilon_r \cdot 2 \frac{K(k_1)}{K(k_1')} + \epsilon_0 \cdot 2 \frac{K(k_0)}{K(k_0')} = 2\epsilon_0(\epsilon_r + 1) \frac{K(k_0)}{K(k_0')}. \quad (3.2.1-8)$$

In [32], the approximations for the ratios of the complete elliptic integrals are given as

$$\frac{K(k)}{K(k')} = \frac{1}{\pi} \ln \left[2 \frac{1 + \sqrt{k}}{1 - \sqrt{k}} \right] \quad 0.707 \leq k \leq 1, \quad (3.2.1-9)$$

$$\frac{K(k)}{K(k')} = \frac{\pi}{\ln \left[2 \frac{1 + \sqrt{k'}}{1 - \sqrt{k'}} \right]} \quad 0 \leq k \leq 0.707.$$

The method treats the structures as a transmission line totally immersed in a homogenous dielectric material with an effective dielectric constant determined by the ratio of capacitance of the line with dielectric material presented to that of the line with the dielectric material replaced by air, so the effective dielectric constant is given as

$$\epsilon_{eff} = \frac{C}{C_{air}} = \frac{1}{2}(\epsilon_r + 1). \quad (3.2.1-10)$$

The characteristic impedance of the transmission line is given as

$$Z_{cpw} = \frac{1}{C \cdot v_{cpw}} = \frac{30\pi}{\sqrt{\epsilon_{eff}}} \frac{K(k_1)}{K(k_1')}, \quad (3.2.1-11)$$

where C is the capacitance per unit length of the transmission line and $v_{cpw} = c/\sqrt{\epsilon_{eff}}$ is the guided wave velocity in CPW line, c is the velocity of light in vacuum.

The above derivation shows us conformal mapping is a simple and fast method to analyze CPW transmission line under the Quasi-TEM assumption. It is shown that data provided by quasi-static approximation deviates from that of rigorous spectral domain hybrid mode approach by less than 1 percent for a CPW line with slot width of 20 μm , and the smaller the slot width, the higher the accuracy [17].

C. Veyres and V. F. Hanna [30] also verified the accuracy of the simple single layered model with the experiment result from Muller at a frequency of 4 GHz, the comparison of effective dielectric constants showed less than 4 % difference for CPW line dimensions used in the experiment.

3.2.2 Multi-layer CBCPW model

In microwave integrated circuits, CPW lines have complex structures and the simple single layered model is not accurate enough for CAD analysis and simulations where multi-layer structure and packaging effect have to be considered. Recent progress in High Temperature Superconductor (HTSC) microwave devices and integrated optoelectronic devices have initiated extensive study of multi-layer CPW structures, which are of practical importance for CAD orientated simulations. This is especially true for GaAs based integrated optoelectronic devices widely used for high-speed

telecommunication applications. GaAs as well as other similar crystal semiconductor materials are in practice used in thin layers and they are fragile, so either dielectric material or metal layer is used as supporting layer underneath the thin semiconductor layer. Besides, integrated optoelectronic circuits inherently involve multi-layered structures where multi-layer optic slab waveguides are often found to be sandwiched in between planar microwave circuits or packaging walls. More sophisticated model has to be used to address these changes and their influence on conventional CPW lines.

Basically, the model for multi-layer CPW structures is based on conformal mapping and partial capacitance techniques [33] under the Quasi-TEM approximation. The partial capacitance technique is based on splitting the multi-layer substrate into several single-layer 'sub-structures' with modified dielectric constants according to the physical layout and line geometry as shown in Fig. 3.2.2.2(a) to Fig. 3.2.2.2(d). Then each of the sub-layers with modified dielectric constant is mapped into a finite rectangular image domain by conformal mapping and Christoffel-Schwartz transformation in a manner similar to that shown in single-layer CPW structure. The partial capacitance per unit length for each of these single-layer 'sub-structures' can be calculated, and the total capacitance per unit length is the sum of all the partial capacitances, and could be used to derive the transmission line parameters. The Veyres-Hanna approximation (superposition of partial capacitances) is extended to this multi-layer CPW lines. The use of the superposition of the partial capacitances is exact when all the boundaries of the dielectrics are along the electric field lines [16]. In this case the magnetic wall can be assumed along the dielectric boundaries and the capacitance of a transmission line can be divided into partial capacitances [34]. In the ideal single-layer

CPW line, this boundary condition is satisfied, but for the multi-layer model, the solution from this method is an approximation, which has been tested by E. Chen and S. Y. Chou [16]. In their work, the calculated parameters of a CPW line on oxide and silicon was used to compute the group velocity and compared with the experimental data. The simulation showed a less than 3% discrepancy from the experiment.

The following example gives an illustration on how the analysis works. Suppose we have a shielded dielectric loaded multi-layer CPW structure as below:

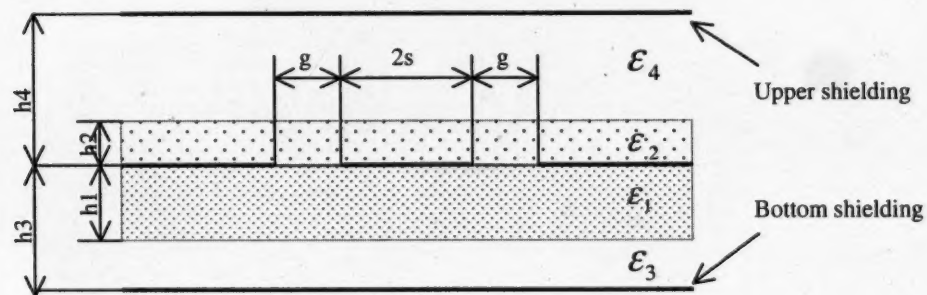


Fig. 3.2.2.1: Two-layer CPW line with upper and bottom shielding [33]

The two dielectric layers on either sides of the CPW line have the dielectric constant of ϵ_1 and ϵ_2 with the thickness of h_1 and h_2 . The upper and lower metal shielding are introduced by either package or supporting metal walls and assumed to be of zero thickness. Two layers of air are assumed to be between the top and bottom metal plates and dielectric material for the sake of simplicity. This model is the basis for CBCPW and loaded CBCPW structures analysis.

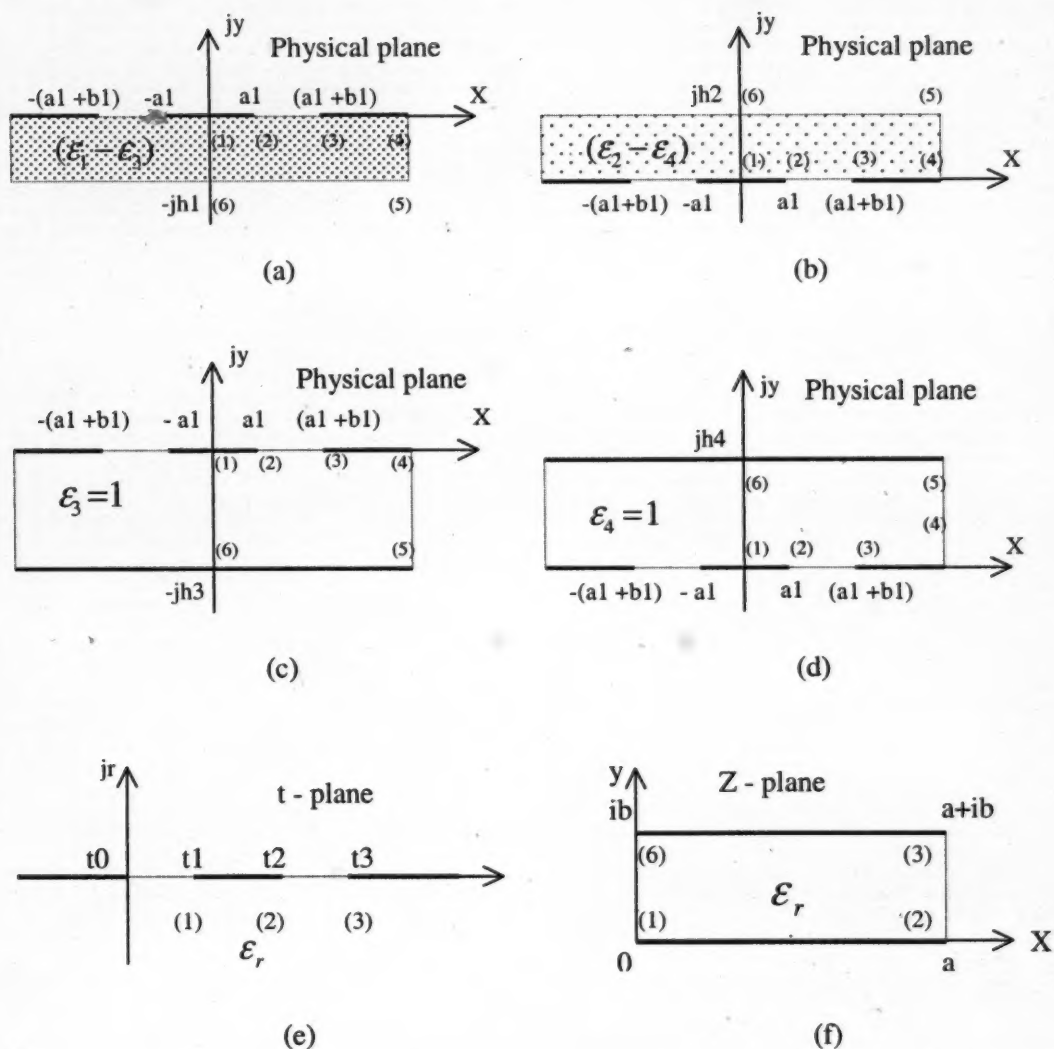


Fig. 3.2.2.2: Conformal transformation of multi-layer CBCPW

First, the two-layer CPW line is decomposed into four 'sub-structures' as shown in Fig.3.2.2.2 (a)-(d). The layers next to metal shieldings are assumed to be air filled, and Fig. 3.2.2.2(c) and Fig. 3.2.2.2(d) represent the transmission line structure in the absence of dielectric layer, the partial capacitance due to these free spaces are denoted by C_{03} and C_{04} . The other partial capacitances denoted as C_1 and C_2 are due to the

dielectric layers next to the CPW line on the bottom and top with equivalent relative permittivity of $(\epsilon_1 - 1)$ and $(\epsilon_2 - 1)$ as shown in Fig. 3.2.2.2 (a) and Fig. 3.2.2.2 (b).

In Fig.3.2.2.2 (c) and Fig.3.2.2.2 (d) of air filled structures, the static capacitances are determined by the dielectric permittivity ϵ_0 of air and the distance between the electrodes in Z-plane. We follow the procedures described in the previous section and formulae (3.1-1) to (3.1-4). After two steps of conformal transformations the partial capacitances due to the upper and bottom substrates are obtained as

$$C_{03} = 2\epsilon_0 \frac{K(k_3)}{K(k'_3)}, \quad C_{04} = 2\epsilon_0 \frac{K(k_4)}{K(k'_4)} \quad (3.2.2-1)$$

where the modulus of the complete elliptic integral, $k_{3,4}$ are

$$k_3 = \frac{\tanh(\frac{\pi s}{2h_3})}{\tanh(\frac{\pi(s+g)}{2h_3})}, \quad k_4 = \frac{\tanh(\frac{\pi s}{2h_4})}{\tanh(\frac{\pi(s+g)}{2h_4})} \quad (3.2.2-2)$$

The use of the hyperbolic tangent function in the transformation is due to the introduction of the conductor shielding, which causes a radical change of electromagnetic fields right above the conductor shielding.

For the two dielectric filled structures in Fig. 3.2.2.2 (a) and Fig. 3.2.2.2 (b), the same conformal mapping procedures are followed, and the static partial capacitances C_1 and C_2 are determined by the equivalent dielectric constants of the two structures and the distance between the electrodes in the Z-plane as shown in Fig. 3.2.2.2 (f)

$$C_1 = 2\epsilon_0(\epsilon_1 - 1) \frac{K(k_1)}{K(k'_1)}, \quad C_2 = 2\epsilon_0(\epsilon_2 - 1) \frac{K(k_2)}{K(k'_2)} \quad (3.2.2-3)$$

The modulus of the complete elliptic integral, $k_{1,2}$ are hyperbolic sine functions

$$k_1 = \frac{\sinh(\frac{\pi \cdot a_1}{2h_1})}{\sinh(\frac{\pi \cdot b_1}{2h_1})} \quad k_2 = \frac{\sinh(\frac{\pi \cdot a_1}{2h_2})}{\sinh(\frac{\pi \cdot b_1}{2h_2})} \quad (3.2.2-4)$$

The total capacitance of the shielded line is:

$$C = C_1 + C_2 + C_{03} + C_{04} = 2\epsilon_0 \left[(\epsilon_1 - 1) \frac{K(k_1)}{K(k_1)} + (\epsilon_2 - 1) \frac{K(k_2)}{K(k_2)} + \frac{K(k_3)}{K(k_3)} + \frac{K(k_4)}{K(k_4)} \right] \quad (3.2.2-5)$$

The effective dielectric constant of the structure is obtained by using (3.2.1-10),

where C_{air} is the line capacitance with all the dielectric materials replaced by air and is given as

$$C_{air} = C_{03} + C_{04} = 2\epsilon_0 \left[\frac{K(k_4)}{K(k_4)} + \frac{K(k_3)}{K(k_3)} \right]$$

So, ϵ_{eff} can be obtained

$$\begin{aligned} \epsilon_{eff} = \frac{C}{C_{air}} &= \frac{(\epsilon_1 - 1) \frac{K(k_1)}{K(k_1)} + (\epsilon_2 - 1) \frac{K(k_2)}{K(k_2)} + \frac{K(k_3)}{K(k_3)} + \frac{K(k_4)}{K(k_4)}}{\frac{K(k_4)}{K(k_4)} + \frac{K(k_3)}{K(k_3)}} \\ &= 1 + (\epsilon_1 - 1) \frac{K(k_1)}{K(k_1)} \left[\frac{K(k_4)}{K(k_4)} + \frac{K(k_3)}{K(k_3)} \right]^{-1} + (\epsilon_2 - 1) \frac{K(k_2)}{K(k_2)} \left[\frac{K(k_4)}{K(k_4)} + \frac{K(k_3)}{K(k_3)} \right]^{-1} \\ &= 1 + q_1(\epsilon_1 - 1) + q_2(\epsilon_2 - 1) \end{aligned} \quad (3.2.2-6)$$

where the coefficients q_n are filling factors to represent the weighted effects introduced by every dielectric layer whose dielectric constant is not equal to unity.

The characteristic impedance of the line is given by equation (3.2.1-11) as

$$Z_{cpw} = \frac{1}{C \cdot v_{cpw}} = \frac{60\pi}{\sqrt{\epsilon_{eff}}} \left[\frac{K(k_3)}{K(k_4)} + \frac{K(k_4)}{K(k_3)} \right]^{-1} \quad (3.2.2-11)$$

Since k_n values can be calculated from the geometry of CPW line as in formula (3.2.2-2) and (3.2.2-4), the line impedance can be obtained.

These models obtained by conformal mapping provide us with fast and simple solutions for CPW and CBCPW analysis. However they are 2D quasi-static models and the frequency dependent dispersion and loss phenomenon are not included, which is why this method can only be used in low microwave frequency range where the frequency dependant effect is small enough to be ignored.

3.3. Losses and attenuation in CPW and CBCPW lines

As mentioned in section 2.3, there are three major loss mechanisms in all transmission lines: conductor loss, dielectric loss and radiation loss.

3.3.1 Conductor loss in CPW line

Conductor loss is presented in all metals and is caused by the finite conductivity σ of the conductor. The conductor loss is also known as ohmic loss or "copper loss", and is frequency dependent. The higher the working frequency the higher the loss due to the skin effect, where current would like to flow on the conductor surface at high frequencies. Since resistance is inversely proportional to the cross-sectional area of the conductor, it increases as the frequency increases. Also, since power loss increases as resistance increases, power losses increase with frequency because of the skin effect. The skin depth is defined as the distance from the metal surface, where the electric field

strength is reduced to 1/e of its peak value. The skin depth is given by $\delta = \sqrt{f \cdot \pi \cdot \mu \cdot \sigma}^{-1}$, and it has a square root dependence on frequency [6][36] [39-41].

Conductor loss evaluation is one of the most difficult problems in transmission line analysis [38], the major problem with accurate prediction of conductor loss lies in the fact of failure to model the transition from low frequency to high frequency. The conventional conductor loss calculation techniques are [42]:

- Simple DC+ skin depth resistance model
- Incremental inductance rule (Wheeler)
- Conformal mapping
- Full wave calculation

The DC + skin depth resistance model states that, at low frequency where the conductor thickness is much smaller than the skin depth, the current distribution in conductor is uniform across the sectional area of the conductor, and the DC resistance is the source of conductor loss (I^2R). As frequency increases, the conductor thickness is sufficiently larger than the skin depth, and the current flow in the conductor tends to concentrate to the surface of the conductor and the real part of the so-called surface impedance becomes the source of the loss [39]. However, this model produces large deviation from the measured result at high frequencies [42].

H. A. Wheeler proposed the famous incremental inductance rule [43]. In this method, the conductor loss is calculated using the complex surface impedance and taking into account the high frequency skin depth effects [44], and it is much more accurate than

the usual perturbation method, where a perfect conductor is assumed and internal inductance is neglected. The conductor loss formula for CPW based on wheeler's incremental inductance rule is given as [15][45]

$$\alpha_c = 4.88 \cdot 10^{-4} \cdot R_s \cdot \epsilon_{eff} \cdot Z_{cpw} \cdot \frac{P'}{g \cdot \pi} \left(1 + \frac{s}{g}\right) \left\{ \frac{\frac{1.25}{\pi} \cdot \ln \frac{4\pi \cdot s}{t} + 1 + \frac{1.25t}{\pi \cdot s}}{\left[2 + \frac{s}{g} - \frac{1.25t}{\pi \cdot g} \left(1 + \ln \frac{4\pi \cdot s}{t}\right)\right]^2} \right\}$$

(dB / unit length),

where R_s is the surface resistance of the metal in CPW structure and is given as

$$R_s = \frac{1}{\sigma \cdot \delta} = \sqrt{\pi \cdot f \cdot \mu_0 / \sigma},$$

ϵ_{eff} and Z_{cpw} are the effective dielectric constant and characteristic impedance of CPW

line, s is the center conductor width, g is the gap width, t is the conductor thickness, and

$$P' = \frac{k}{(1-k')(k')^{\frac{3}{2}}} \left[\frac{K(k)}{K(k')} \right]^2 \quad \text{for } 0 \leq k \leq 0.707$$

$$= \frac{1}{(1-k)\sqrt{k}} \quad \text{for } 0.707 \leq k \leq 1.0$$

where k and $K(k)/K(k')$ are defined as in equation (3.2.2-4) and (3.2.1-9) respectively,

and $k' = \sqrt{1-k^2}$.

The simulation of conductor loss using this formula closely matches the measured transmission loss curve of fabricated CPW line on quartz up to 20 GHz indicating conductor loss is the main loss mechanism in CPW lines in this frequency range. The

simulated conductor loss and measured loss curves in CPW of the same length are shown in Fig. 5.1.4.

Conformal mapping techniques can also be used to calculate conductor loss associated with skin depth and current crowding [38]. The frequency dependence of the line parameters is included by introducing a frequency term. The simulated results showed excellent agreement with their measurements and experiments result from other workers [46]. However, no detailed information or closed form formula for calculating the attenuation constant was given.

Wolfgang Heinrich gave a quasi-static description of CPW transmission line and closed form solutions of the distributed parameters and conductor loss, which include the effect of non-ideal conductors and finite metallization thickness [47]. However the formulae are fairly complicated and the derivation involves curve-fitting technique. He also conducted full wave analysis employing mode-matching technique for conductor loss in lossy MMIC transmission lines [48]. His study showed that CPW lines are quite sensitive to metallization thickness, which should be at least 3 times the skin depth.

Rastogi et al. [40][49] observed that the conductor loss increases with the increase of relative dielectric constant of CPW while the dielectric loss depends mainly on the substrate thickness and affected little by the line dimensions. A guideline on conductor metallization thickness of 3 μm is given in their work as a reasonable estimation for practical microwave applications for any conductors at frequency above 3 GHz. Their simulation showed that the higher the conductivity the lower the conductor loss and it becomes more prominent as frequency increases. However, since the skin

resistance decreases only as the inverse of the square root of the conductivity of the conductor, the increase of conductivity in practice will not lead to much reduction in conductor loss [50].

Chun-Lin Liao and his co-workers [51] identified the dominant loss mechanism as conductor loss at low microwave frequencies in the CPW structure investigated. Similar result was obtained by Jeng-Yi Ke and Chun Hsiung Chen [52]. They also found that the smaller the line dimension, the more sensitive the conductor loss is to the metallization thickness.

In summary, the conductor loss could be simulated by quasi-static models; however these models either involve empirical formulae and geometry dependant constants or could not give closed form solutions, thus their usage is limited. The conductor loss curve produced by incremental inductance rule for up to 20 GHz closely matches the measured transmission loss in fabricated CPW line on quartz indicating it is a valid and fairly accurate model for predicting conductor loss in the investigated frequency range.

3.3.2 Dielectric loss and dispersion in CPW line

The dielectric loss and dispersion on the other hand is determined by the characteristics of dielectric materials, which are usually classified by their electrical breakdown strength, dielectric loss, permittivity, and polarization [35]. In understanding the dispersion and dielectric loss effects, polarization is the key linking the macroscopic properties and the atomic-molecular theory of dielectric materials. Microscopically, dielectric polarization occurs when the positive and negative charges experience an

electric force that causes them to move apart in the direction of the external field. As a result the centers of positive and negative charges no longer coincide and the material obtains a certain polarity. Three types of polarizations are mainly involved, which are electronic, ionic and dipole orientation. The mechanisms of polarization have varying time response capability to an applied field frequency, and the net contribution of polarization to the dielectric constant is therefore frequency dependant. Since the superposition of the polarizations determines the dielectric constant of a material, the dielectric constant is expected to change over a wide frequency range due to the change in the total polarization. This directly leads to the dielectric dispersion effect.

Most of the dielectric materials are classified as insulators, because there are virtually very few free moving charges, and the dielectric character is mainly determined by the polarizations of bound charges in the material, so the conduction loss in dielectric materials due to finite conductivity can be largely ignored. For a static situation, the polarization is in phase with the external electric field, and there is no dielectric loss. However, in the presence of an alternating electric field, the polarization vector of the material lags behind the applied electric field leading to energy dissipation in forms of oscillation and damping process in the material on the molecular scale [36]. Depending on the ability to respond to an electric field of different polarization mechanisms, dielectric loss varies at different frequencies. Electronic and ionic polarizations usually happen at high frequencies (> 1000 GHz), and dipole orientation is the major cause of dielectric loss in polymer materials at much lower frequencies [65]. The dielectric loss is measured by loss tangent, the bigger the value the larger the dielectric loss. The dielectric loss in CPW lines is relatively independent of line geometry [47], but

determined by the property of dielectric substrate, such as loss tangent and intrinsic resistivity of the material. For semiconductors with high resistivity and low loss tangent less than 0.001[22], such as GaAs, Quartz and most organic polymeric materials, the dielectric loss can be ignored compared to the conductor loss in a transmission line.

3.3.3 Radiation loss in CPW line

Radiation loss will occur in open structure or any discontinuities in transmission lines, such as steps, bends, or any abrupt geometric changes related to impedance changes. In real transmission lines, discontinuities are inevitably involved, and they are the sources of electromagnetic disturbances and radiations. The transmission and radiation of electric energy in planar lines is shown in ray diagram in Fig.3.3.3.1

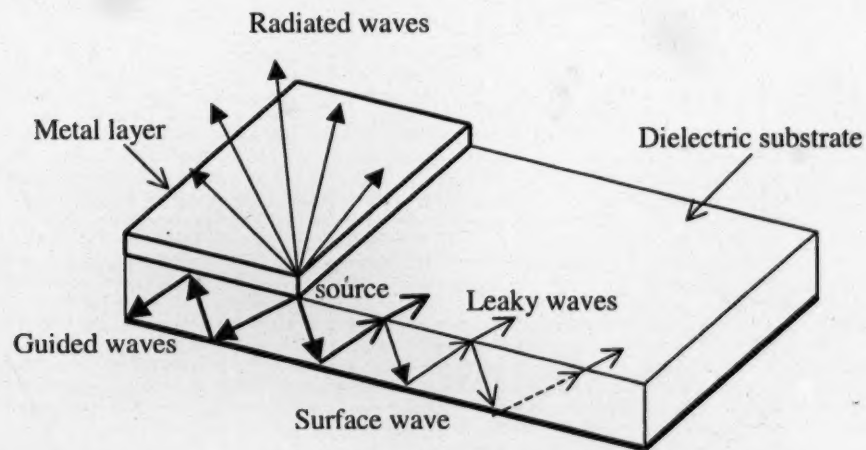


Fig. 3.3.3.1: Four kinds of waves emitted by a point source

It is assumed that a point current source (Hertz dipole) located on the edge of the upper conductor and radiates electromagnetic waves in all directions [37]. Depending on the angles toward which they are transmitted, the waves can be categorized as guided

wave that contributes to signal power transmission, leaky and surface waves that contribute to losses in the transmission line.

- Guided waves

These waves located between the top and bottom conductors and are guided within substrate by reflection on the two metal walls.

- Radiated waves

These waves propagate upwards away from the metal conductor, and produce leakage and are not wanted in open transmission line design.

- Leaky waves

These waves propagate in the substrate and are reflected by the lower ground plane to the air-dielectric interface. Depending on the angle of incidence some waves are partly reflected and partly transmitted, they are called leaky waves and they contribute to radiation loss.

- Surface waves

Waves with larger incident angles in the substrate are totally reflected from the air-dielectric interface and propagate with small attenuation. They are called surface waves and can produce spurious coupling between circuits.

In CPW lines, radiation loss happens at high frequencies where mode conversion happens, energy of CPW mode would leak into other unbounded modes and radiated away from the transmission line. Many workers have had valuable contributions to the investigation of radiation loss in CPW lines.

In 1990, H. Shaigesawa et al. [53] identified for the first time the mode coupling effect between the standard dominant CPW mode and other surface-wave-like leaky modes shown in Fig. 3.3.1 above a certain critical frequency, when the bonded fundamental CPW mode begins to leak energy into higher order surface-wave-like leaky mode and radiate away [27] [54].

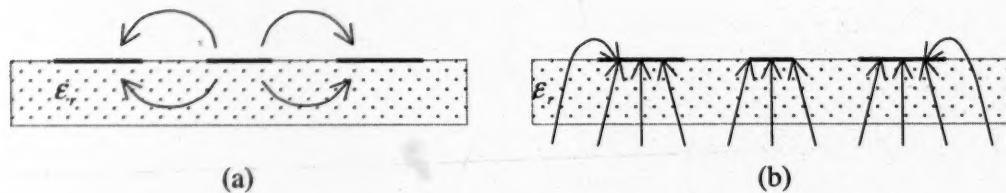


Fig. 3.3.3.2: (a) Standard CPW mode (b) Surface-wave-like leaky mode [53]

They also investigated the structural effect on leaky behavior in finite width CPW lines [55], and observed the attenuation caused by leaky modes depends very strongly on the ratio of side conductor width to the substrate thickness while rather weakly on the ratio of gap width to the substrate thickness. Riaziat et al. [56] conducted similar study and found out that with the increase in frequency, more leaky modes interact with CPW mode causing bigger attenuation in CPW lines. To avoid the leaky surface-wave modes, the substrate thickness has to be smaller than $0.12 \cdot \lambda_g$, where λ_g is the guided wavelength.

Gopalakrishnan et al. [57] also investigated the electrical losses in CPW structures, and attributed the radiation loss to the phase-matched mode coupling between the CPW mode and off-axis substrate modes. The critical frequency at which the mode coupling effect takes place is closely related to the thickness of the substrate [58], the

thinner the substrate, the higher the critical frequency would be. Their result shows a critical frequency about 25 GHz for a CPW line on 500 μm thick substrate with a microwave index near 2.4, which resembles our CPW on quartz structure. This indicates that we might be able to avoid radiation loss associated with mode coupling effect in our CPW line at 20 GHz.

In summary, conductor loss, dielectric loss and radiation loss contribute to signal attenuation in transmission lines. These losses are more or less frequency dependent and give rise to dispersion effect, which leads to amplitude and phase distortion. These frequency dependent loss properties are usually ignored at low frequency, but they become more significant in high-speed digital systems. When a high-speed digital signal is transmitted through a lossy transmission line, the sharp edge becomes rounded due to a bigger loss in high frequency components.

Conductor loss due to the finite conductivity is dominant as the frequency is less than the critical frequency where the leakage or radiation loss becomes bigger than conductor loss. For CPW lines fabricated in this project, the radiation loss is not a concern in the investigated frequency range where the leaky modes are not excited [56].

3.3.4 Parallel-plate power leakage loss in CBCPW line

In CBCPW structures, besides the three types of losses mentioned above, the biggest concern is the possible excitation of parallel-plate leakage modes loss due to the conductor-backed configuration [8]. The power leakage from the dominant CPW mode to the leaky modes can be particularly severe for CBCPW lines with thin dielectric layer

at high frequencies. The electric field configurations of the CPW mode and some of the parallel-plate leaky modes are shown in Fig. 3.3.2 [59].

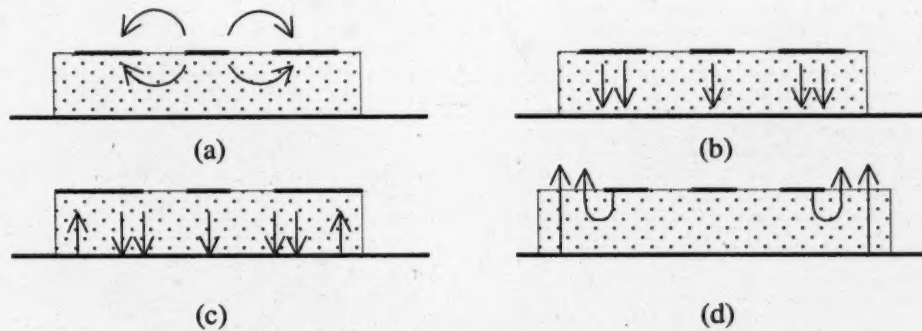


Fig. 3.3.3.3: (a) CPW mode (b) Micro-strip mode (c) Image guide like mode (d) First higher order micro-strip mode

The CPW mode and the micro-strip mode are two dominant propagation modes below a certain critical frequency where other higher order leaky modes can be activated, and mode conversions among the incident, transmitted and reflected waves take place. The over-mode nature of CBCPW structure explains the energy-leaking problem. Finite electrical potentials could appear on the side ground conductors due to the difference in phase velocities of different propagation modes, and electric energy is carried away by these varying electrical potentials on the side conductors if they are not grounded along their whole lengths. However, those higher order effects should be minor in the frequency range where Quasi-TEM approximation is valid.

C. Tien et al. [59] gave a very comprehensive review on the transmission character of CBCPW lines of finite width. In their experiment, the periodic resonant-like transmission character measured in the spectrum domain was identified as related to the micro-strip mode shown in Fig. 3.3.2(b). They also showed a successful suppression of

the micro-strip mode by uniformly grounding the side ground conductors with the backside conductor along the whole length of the transmission line up to 17 GHz, but as frequency goes higher, first higher order micro-strip mode was excited and contribute to transmission loss.

M. Riazat et al. [60] pointed out that the CPW mode actually is not a normal propagation mode in a CBCPW structure. The normal propagation modes are so called 'slot line mode', 'micro-strip mode' and 'coplanar' mode, and the desired CPW mode is a superposition of 'micro-strip' mode and 'coplanar' mode. Since the phase velocities of these two modes are different, the phase angle between them drift along the line and as a result, a finite potential appears on the side ground conductors, which gives rise to parallel plate leaky modes. One fact we can learn from their results is that in finite CBCPW transmission lines, it is impossible to remove those higher order surface leaky modes unless we can properly ground the side conductors, otherwise there would be finite potentials on them causing leaking problems. The proper grounding can be achieved by vias or wrap-around technology, which however increases the fabrication difficulties. These higher order effects are non-negligible at high microwave frequencies, but the severity of this effect at the low microwave frequency of our interest is yet to be investigated.

3.4. Methods to suppress the power leakage in CBCPW

The CPW mode is subject to the influence of the lower ground conductor layer especially when the substrate thickness is comparable to the slot width between center conductor and the side ground planes [59] giving power loss into higher order leaky

modes in the structure. N. K. Das provided us with insightful analysis and methods of solving the problem [61]. The most convenient and economical methods are either using a two-layer CBCPW configuration or using dielectric loading on top, which would make the propagation constant of the CPW mode bigger than that of other higher order modes and suppress the leakage. The cross section view of the configurations are shown below:

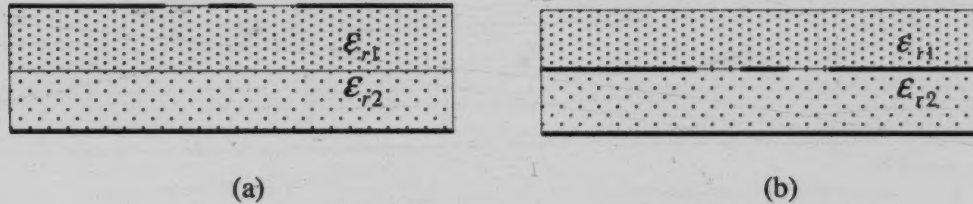


Fig. 3.4: (a) Two-layer CBCPW (b) Dielectric loaded CBCPW

Two dielectric materials with dielectric constants ϵ_{r1} and ϵ_{r2} ($\epsilon_{r1} > \epsilon_{r2}$) are used in the CBCPW structures. By careful selection of the dielectric constants of the substrate material and proper arrangement, we can satisfy the requirement that the propagation speed of CPW mode is slower than the parallel plate mode and the power leakage is suppressed.

Multi-dielectric-layer structures are also used to suppress the geometry and frequency related power leakage problem by Itoh and his coworkers [62]. By using spectrum domain analysis they showed that whenever the phase constant of the dominant CPW mode is bigger than that of the parallel plate leaky mode, the attenuation constant is greatly reduced.

Comparing these two methods, the dielectric loaded structure is easier to fabricate. Because the top CPW metal layer deposition, which usually involves high

temperature process or chemical environment could be done before the coating of the top dielectric layer, which minimize the influence of metal deposition process imposed on the dielectric material.

4. Experiment and measurement

4.1. Measurement basics and S-parameter definition

Vector Network Analyzer (VNA), a microwave measurement system, is commonly used to measure the reflection and transmission coefficients or scattering parameters (S-parameter). It gives the information of the return loss, which is the ratio of the reflected power to the incident power (reflection loss) and the transmission loss, which is the ratio of the transmitted power to the incident power (attenuation) in microwave frequency range. Some terminologies and conventions in microwave measurement are explained as well as the two transmission loss measurement methods in this section.

- **S-Parameter definition and conventions**

The diagram of S-Parameter measurement for a two-port network is shown in Fig. 4.1. The transmission line is connected to a VNA through its two ports, and it is treated as a two-port network by the VNA. Each port has two nodes: one representing the entering or 'a' wave and one for the emerging 'b' wave. Lines that connect nodes are called branches. Each branch has an arrow and value corresponding to a S-Parameter. Energy will flow only in the direction of an arrow. The S-Parameter numbering convention is S_{outin} , where the first number (out) refers to the port where energy emerges and the second number (in) names the port at which energy enters the network. For example, the S-Parameter S_{21} identifies the measurement as the complex ratio of the emerging wave at the output port with respect to the incident wave at the input port. The S-Parameters in the two-port network are defined by equation (4.1-1).

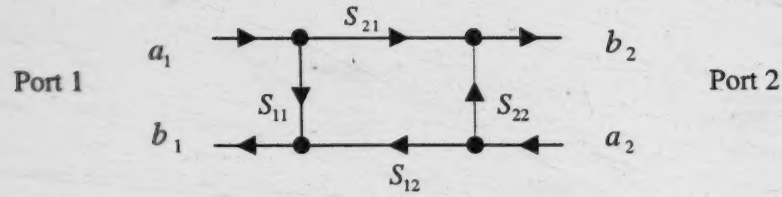


Fig. 4.1: S-parameter diagram

$$\begin{aligned} b_1 &= S_{11}a_1 + S_{12}a_2 \\ b_2 &= S_{21}a_1 + S_{22}a_2 \end{aligned} \quad (4.1-1)$$

a_1, a_2 represent incident electric waves at the two ports of the network, and b_1, b_2 represent electric waves leaving the two-port network. The square of these parameters has the dimension of power, and can be thought of as incident and reflected power waves. The four S parameters are evaluated in terms of the defining equations.

$$S_{11} = \left. \frac{b_1}{a_1} \right|_{a_2=0} \quad S_{21} = \left. \frac{b_2}{a_1} \right|_{a_2=0} \quad S_{12} = \left. \frac{b_1}{a_2} \right|_{a_1=0} \quad S_{22} = \left. \frac{b_2}{a_2} \right|_{a_1=0} \quad (4.1-2)$$

S_{11} is the electric wave leaving port-1 divided by the electric wave entering port-1 when there is no signal enters port-2, and it is also called reflection coefficient representing return loss. The $a_2 = 0$ condition is satisfied if the two-port network has a matched load (usually 50Ω) connected at port-2, so there is no reflected signal coming back into the network. S_{21} is the electric field leaving port-2 divided by the electric field entering port-1 when there is no signal enters port-2, and it is also called transmission coefficient representing transmission loss. Since a transmission line is a passive component and could be considered as a reciprocal network, the input and output port could be switched freely, S_{22} and S_{12} have the same definition as S_{11} and S_{21} except in the opposite direction.

Insertion loss measures the reduction of a microwave signal as it passes through a component, which includes not only internal losses such as conductor loss and dielectric loss, but also losses caused by reflection at discontinuities. The term insertion loss is often used incorrectly to mean attenuation, which is the power loss inside the component or internal loss.

The fact that the device under test is embedded in a characteristic impedance of a measurement system (usually, $Z_0 = 50 \Omega$) is fundamental to definitions of S-parameter measurements and an impedance match is required as shown in equation (4.1-2). So, in order to get a real reading of attenuation in the transmission line, we have to remove the return loss by matching the impedance of the transmission line to that of the VNA system in the first place.

4.2. Measurement method and device design

The loss in CBCPW line is composed of conductor loss, dielectric loss and radiation loss. However based on the previous analysis and literature review, we estimated that the radiation loss is not a concern in the frequency range of interest (up to 20 GHz). The same assumption was made about the dielectric loss, due to the low dispersion and low loss nature of BCB polymer in this frequency range [10]. The major task was to design an effective method to differentiate and identify the parallel-plate leakage loss in CBCPW line from the total transmission loss. The matched-through-line method and resonance method were used to investigate the transmission loss and dispersion effect in CBCPW line in the experiment.

4.2.1. Matched-through-line method

In this method, transmission loss can be directly measured from impedance matched $50\ \Omega$ lines. Three types of transmission lines were needed in order to identify the parallel-plate leaky mode loss. These are $50\ \Omega$ CWP lines on quartz wafer, $50\ \Omega$ CPW lines on BCB layer on quartz wafer, and $50\ \Omega$ CBCPW line with BCB as dielectric material. $50\ \Omega$ dielectric loaded CBCPW lines were also designed to see if the dielectric loading could suppress the parallel-plate leaky mode.

First, CPW lines with $50\ \Omega$ impedance had to be built on low loss substrate to exclude dielectric loss, and identify the pure conductor loss in the line. Quartz, which has a very low dielectric loss, is chosen as the substrate for this purpose. The conductor loss value was obtained by measuring the transmission loss or S_{21} parameter from the VNA.

Second, $50\ \Omega$ CPW lines were built on BCB coated quartz wafer. The transmission loss of such lines should include both conductor loss and dielectric loss if the BCB dielectric layer introduces a measurable dielectric loss. By comparing the results from these two types of CPW lines, it should be possible to determine the effect of the BCB coating in terms of its dielectric loss.

Third, $50\ \Omega$ CBCPW lines with BCB as the dielectric material were built, and transmission loss was measured. As discussed in the previous chapters, a much bigger transmission loss is expected if there are parallel-plate leaky modes. If a bigger than tolerable transmission loss presented in the CBCPW, $50\ \Omega$ dielectric loaded CBCPW line

will be built with a dielectric loading material of certain thickness. If the theoretical analysis is valid, the transmission loss will be reduced in this dielectric loaded CBCPW.

4.2.1.1 Validity of the quasi-static model in transmission line design

50 Ω CPW and CBCPW lines were designed using the quasi-static single-layer CPW model and multi-layer model with metal shielding in sections 3.2. Matlab™ codes were written to simulate the line impedance and effective dielectric constant.

The conductor thickness is assumed to be zero in the models analyzed. However the finite conductor thickness of practical lines give rise to fringing electric field that effectively widens the conductors, and affects the line impedance and effective dielectric constant [15][45]. An investigation was done to determine if the effect should be included in the model by comparing the line impedances obtained by models with and without the effect on a 50 Ω CPW through line calibration standard. The precise 50 Ω CPW through line standard (Dimension: centre conductor width $2s=51.5\text{ }\mu\text{m}$, gap width $g=22.7\text{ }\mu\text{m}$, strip thickness $t=3.5\text{ }\mu\text{m}$, 640 μm alumina substrate with $\epsilon_r=10$) is built on the standard calibration substrate CS-5 by GGB industries Inc. It is designed to precisely calibrate coplanar probes (GSG-150DP) for CPW measurement. The simulation result for a CPW line with the above mentioned dimensions shows excellent agreement with 50 Ω characteristic impedance without including the conductor widening effect. The simulation result including the conductor widening effect shows a large reduction in the impedance from the measured 50 Ω for the through line standard as shown in Fig. 4.2.1.1. By further investigation, we found that this conductor thickness effect model was studied in cases where the ratio of conductor thickness, t , to the gap width, g , is below 0.05. For

the device in our experiment, this ratio is more than 3 times larger. This observation and the large discrepancy between the simulated impedance with the conductor widening effect and the measured impedance indicate that the conductor widening effect should not be included in the simulation model for the device in our experiment.

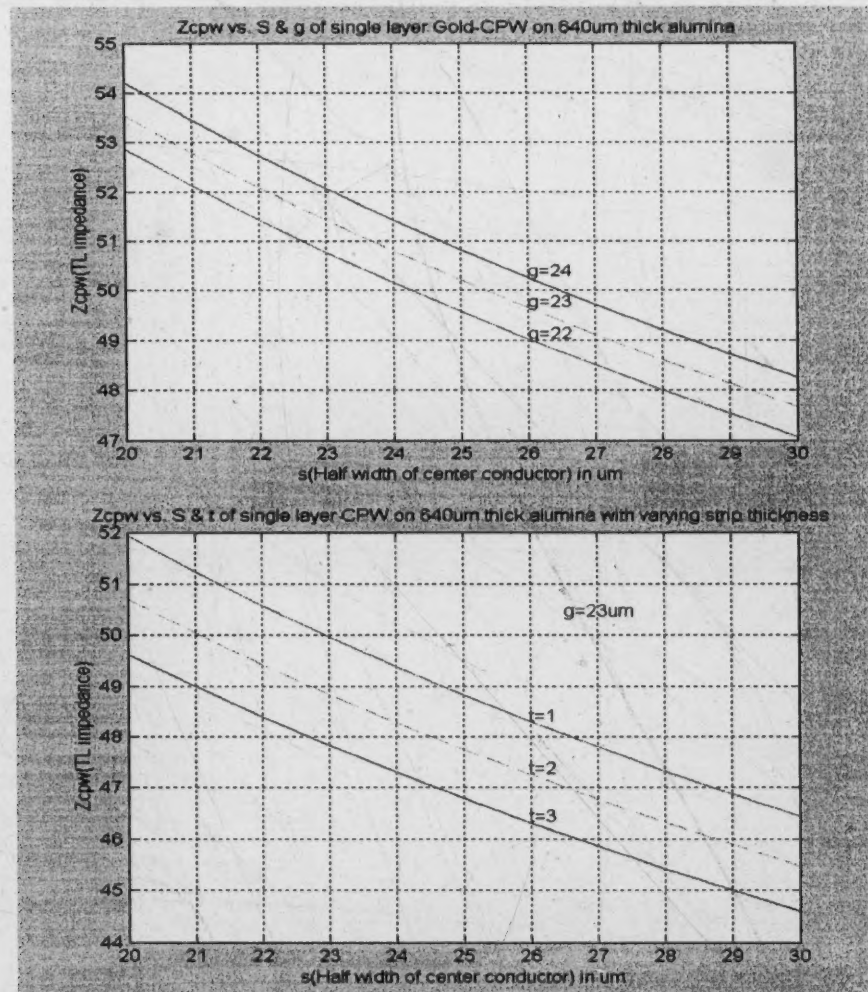


Fig. 4.2.1.1: Simulation results of CPW line impedance for the through line calibration standard with and without conductor thickness effect (g is the gap width in μm , t is the conductor thickness in μm)

Based on this argument, all the CPW and CBCPW lines are simulated with models without including the conductor thickness effect.

4.2.1.2 50 Ω CPW and CBCPW dimensions

Dimensions of 50 Ω CPW on quartz, CPW on BCB on quartz, CBCPW and loaded CBCPW were simulated by quasi-static models. Physical lines were fabricated to compare and investigate the loss mechanisms in them. The critical parameters in determining CPW line impedance are center conductor width, $2s$, and gap width, g . Considering the tolerance in CPW parameters such as substrate dielectric constant and dielectric layer thickness, lines with small difference in gap width were designed and fabricated for each type of these lines in order to achieve characteristic impedance as close to 50 Ω as possible. Fig. 4.2.1.2 shows the cross section views of these lines:

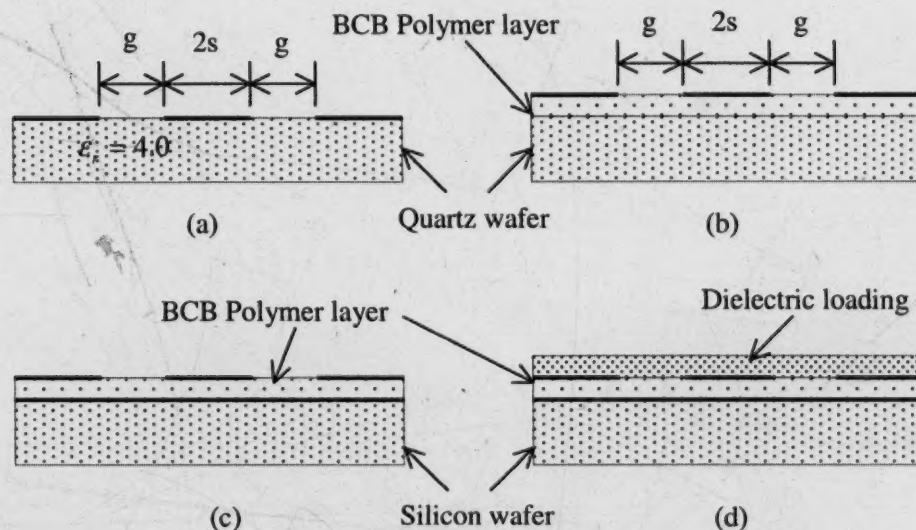


Fig. 4.2.1.2: (a) CPW line (b) CPW line on BCB layer on quartz wafer
(c) CBCPW line (d) Loaded CBCPW line

The following simulation plots and tables show the critical dimensions, effective dielectric constants, and characteristic impedances of these lines:

- CPW on 580 μm thick plain quartz wafer ($\epsilon_r = 4.0$)

Center conductor width $2s$ (μm)	Gap width g (μm)	Effective dielectric constant (ϵ_{eff})	Line impedance Z_{cpw} (Ω)
106	16	2.4961	54.9
106	12	2.4965	51.1
106	10	2.4968	48.9
106	8	2.4971	46.4

Table 4.2.1.2 (a): 50 Ω CPW line design parameter

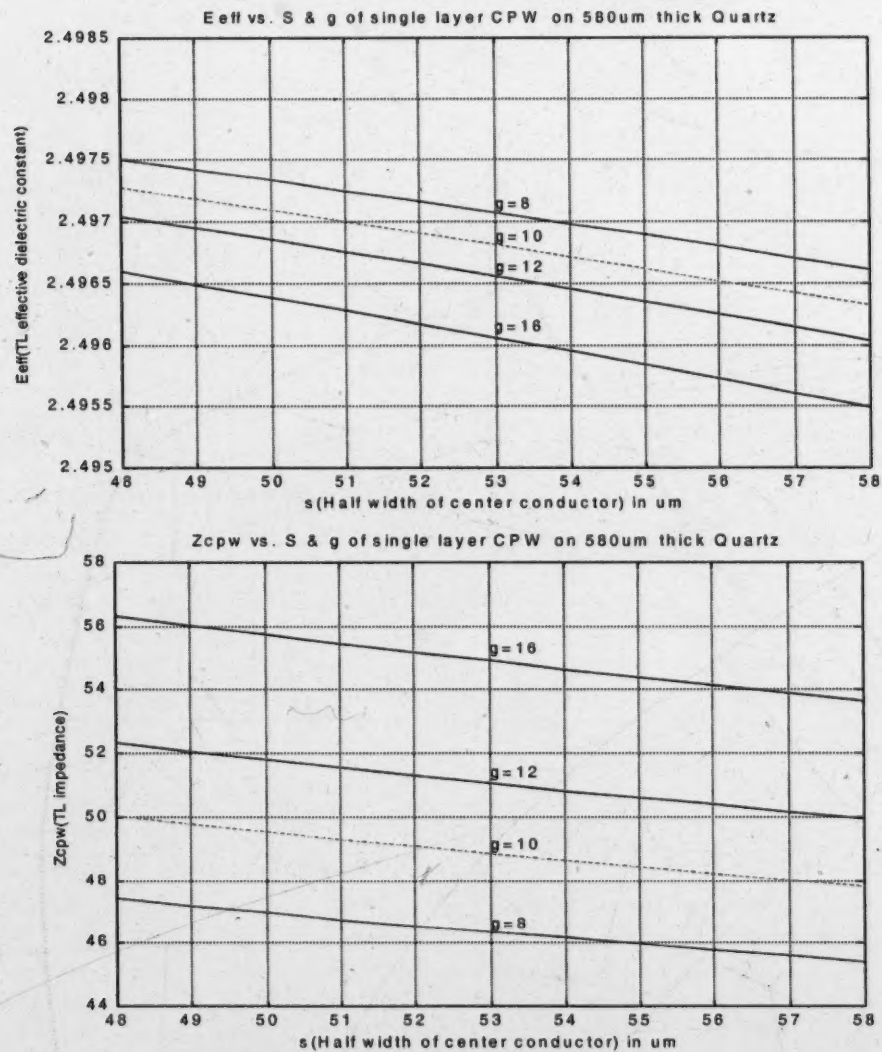


Fig. 4.2.1.3 (a): Characteristic impedance and effective dielectric constant of CPW line with different gap width

- CPW on 18 μm BCB polymer layer ($\epsilon_r = 2.65$) on 580 μm thick quartz wafer

Center conductor width $2s$ (μm)	Gap width g (μm)	Effective dielectric constant (ϵ_{eff})	Line impedance Z_{cpw} (Ω)
146	16	2.1685	54.4
146	12	2.1445	51.0
146	10	2.1306	49.1
146	8	2.1148	46.9

Table 4.2.1.2 (b): 50 Ω CPW line on BCB on quartz design parameters

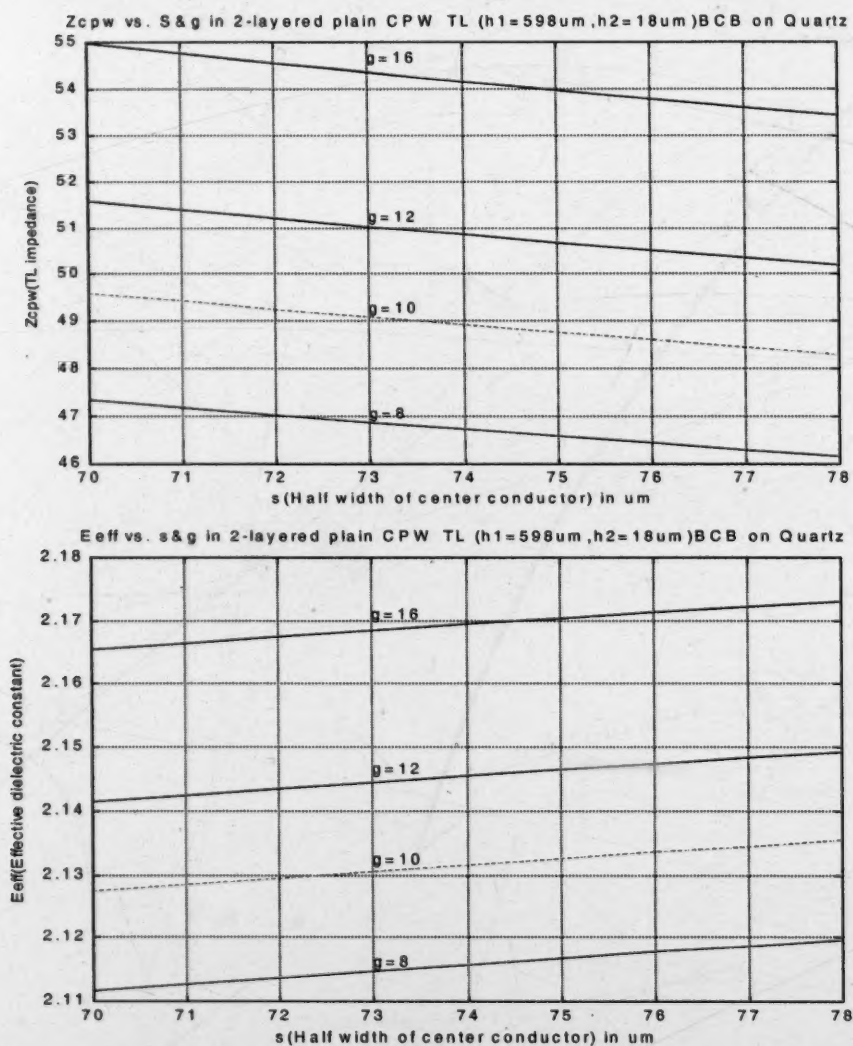


Fig. 4.2.1.3 (b): Characteristic impedance and effective dielectric constant of CPW on BCB and quartz wafer with different gap width

- CBCPW on 18 μm BCB layer ($\epsilon_r=2.65$) on a 500 nm thick conductor layer

Center conductor width $2s$ (μm)	Gap width g (μm)	Effective dielectric constant (ϵ_{eff})	Line impedance Z_{cpw} (Ω)
42	16	2.0862	52.8
42	12	2.0609	51.2
42	9	2.0383	49.4
42	7	2.0210	47.7

Table 4.2.1.2 (c): 50 Ω CBCPW line design parameters

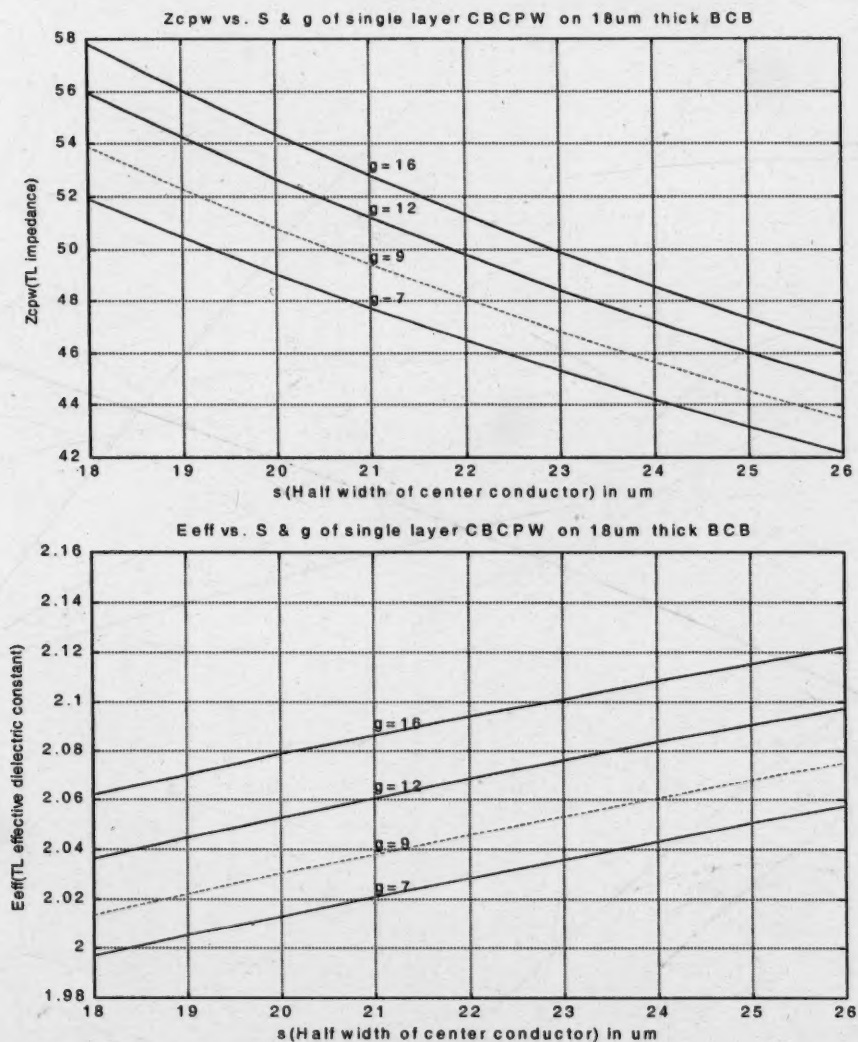


Fig. 4.2.1.3 (c): Characteristic impedance and effective dielectric constant of CBCPW with different gap width

- Loaded CBCPW with a dielectric material of 22 μm thick (assumed $\epsilon_r = 6.6$)

Center conductor width $2s$ (μm)	Gap width g (μm)	Effective dielectric constant (ϵ_{eff})	Line impedance Z_{cpw} (Ω)
22	16	3.8199	54.4
22	12	3.9434	51.2
22	10	4.0111	49.2
22	8	4.0831	46.7

Table 4.2.1.2 (d): 50 Ω loaded CBCPW line design parameter

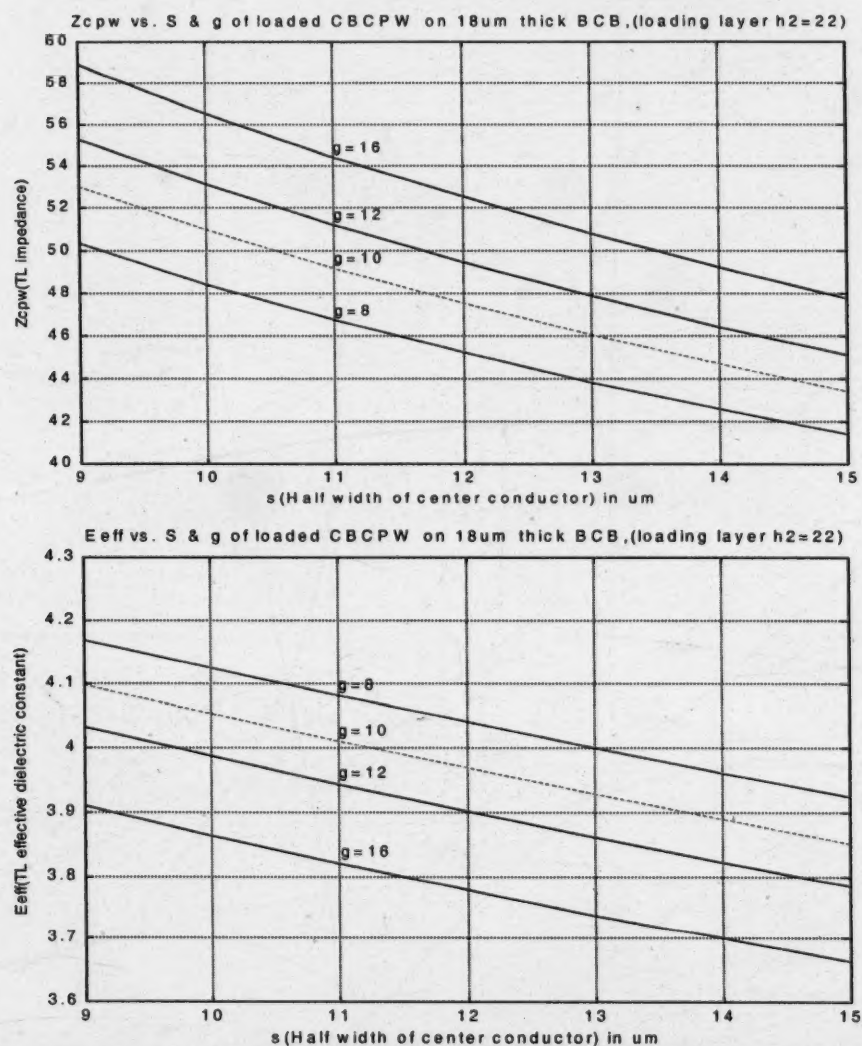


Fig. 4.2.1.3 (d): Characteristic impedance and effective dielectric constant of loaded CBCPW with different gap width

4.2.2. Resonance method

The effective dielectric constant and the dispersion characteristics of CBCPW lines are very important design parameters for EO modulators in order to evaluate the modulation efficiency of the device in the desired frequency range. The dispersion characteristic of a transmission line is related to its geometry and the dielectric material used. The small cross-sectional dimension of CBCPW line has an inherent low dispersion character compared with microstrip line [48][61]. The overall dispersion characteristic of CBCPW is determined by the dispersion characteristic of BCB 3022-63 polymer material used as dielectric layer in the line. The static effective dielectric constant of the designed CBCPW line was obtained using a quasi-static simulation given the information of line geometry and dielectric material. The broadband effective dielectric constant, dispersion character and transmission loss of CBCPW line are experimentally studied by resonance technique. Quarter-wavelength and half-wavelength CBCPW resonators are built for this purpose.

The transmission line resonators are built by introducing discontinuity in the lines such as open gaps and shorted ends. The layout of a quarter-wavelength and a half-wavelength CBCPW resonator are shown in the following figure in a top view:

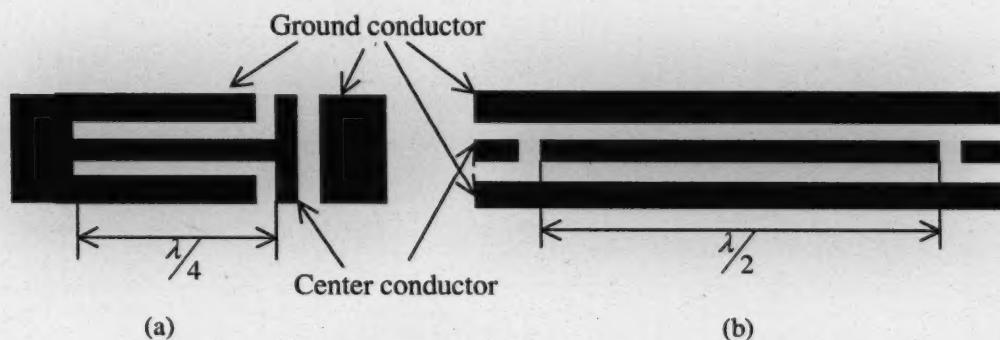


Fig. 4.2.2: (a) Quarter-wavelength Resonator (b) Half-wavelength Resonator

In the quarter-wavelength resonator shown in Fig. 4.2.2 (a), the resonance effect is created by a quarter-wavelength long resonant stub with a shorted or an open end. In our experiment, a shorted resonant stub is used because less radiation loss is generated at the end of the stub compared with an open-end resonant stub. The short-circuited stub behaves like a parallel resonant circuit at frequencies where it is close to an odd multiple of a quarter wavelength ($f_n = (2n + 1)c/4l$), where c is the speed of light, l is the effective length of the resonator and n is an integer), and it is said to be anti-resonant [63]. At these frequencies, the short-circuited resonant stub is seen as an open by the resonator, and maximum power transmission occurs. While in the vicinity of the frequencies where the stub is a multiple of half wavelength ($f_n = nc/2l$), it behaves as a series resonant circuit (Short circuit), where resonance happens with minimum power transmission. So the transmission character of a quarter-wave stub resonator presents transmission peaks at anti-resonant frequencies and dips at resonant frequencies alternatively. In the half-wavelength resonator in Fig. 4.2.2 (b), the discontinuities, made of open gaps, set up resonant standing waves in the center strip. This occurs at frequencies where the effective length of the center line, separated by the gaps, is exactly a half wavelength and a constructive interference between the incident and reflected waves is setup resembling that in a Fabry-Perot resonant cavity.

Half-wavelength and quarter-wavelength resonators were built using CPW, CPW on BCB and quartz, and CBCPW transmission lines. The lengths of the resonators are designed as a half and a quarter of the guided wavelength at 3 GHz based on the estimated static effective dielectric constants.

The loaded Q factors (Q_L), defined as the ratio of center frequency to the half power bandwidth at the vicinity of resonant frequencies, were measured from the CBCPW quarter-wavelength resonator's S_{21} parameter and used to derive the attenuation constant and effective dielectric constant of CBCPW line.

The unloaded Q factor relates to the total transmission loss, and can be obtained by [64]:

$$Q_0 = \frac{Q_L}{\sqrt{1 - 2|S_{21\text{res}}|^2}} \quad (4.2.2-1)$$

where $S_{21\text{res}}$ is the amplitude of measured transmission loss at resonant frequencies.

The effective dielectric constants at the resonant frequencies are given as [64]:

$$\epsilon_{\text{eff},\text{res}} = \left(\frac{n \cdot c}{4 \cdot f_{\text{res}} \cdot l} \right)^2 \quad n = 2, 4, 6, \dots \quad (4.2.2-2)$$

where c is the speed of light in free space, l is the physical length of the resonant stub, and n is even multiple number of half-wavelength where the resonance happen.

The attenuation constant of CBCPW line is obtained by [64]:

$$\alpha_0 = \frac{8.686 \cdot \pi \cdot f_{\text{res}} \cdot \sqrt{\epsilon_{\text{eff},\text{res}}}}{c \cdot Q_0} \quad (\text{dB/cm}) \quad (4.2.2-3)$$

The frequency responses of two identical CBCPW quarter-wavelength resonators were measured. The plots and derived dielectric constants and attenuation constants at descreet resonant frequencies by using the above equations are presented in section 5.

4.3. Mask layout

The four types of $50\ \Omega$ transmission lines, the quarter-wavelength and half-wavelength resonators are laid out on the same mask. Each type of line and resonator share the same critical features and are designed to be fabricated using different substrates. The mask is divided into four sections and each section contains one type of transmission line and resonator as shown in Fig. 4.3.

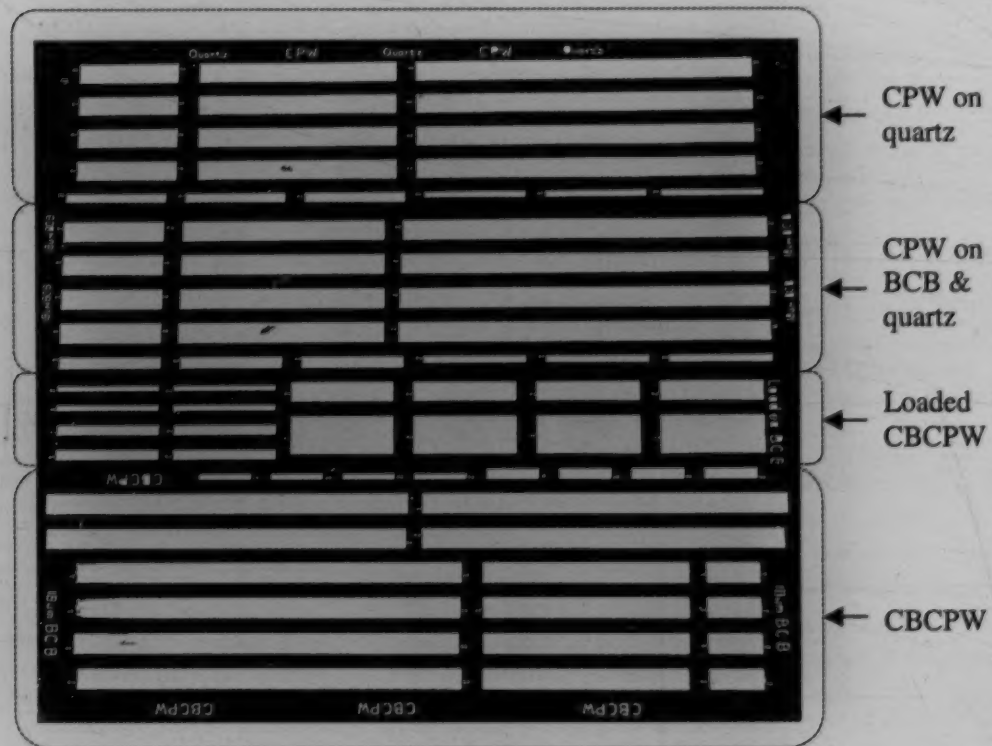


Fig. 4.3: Mask layout

The CPW on quartz section, CPW on BCB & quartz section and CBCPW section consist of 1 cm long through lines, quarter-wavelength resonators and half-wavelength resonators with slightly different gap widths, g . The CBCPW section is composed of 5 mm long through lines, quarter-wavelength and half-wavelength resonators plus four

3.4617 cm long through lines and 5 mm long through lines with different gap width, g. In the loaded CBCPW section, all the lines are through lines with the same length but different gap width and ground conductor width.

4.4. Transmission line fabrication

The CPW lines fabrication process were conducted in a clean room environment in the MicroFab facility on the University of Alberta campus. The process flow chart is shown in Fig. 4.4. Detailed fabrication description is given in the appendix.

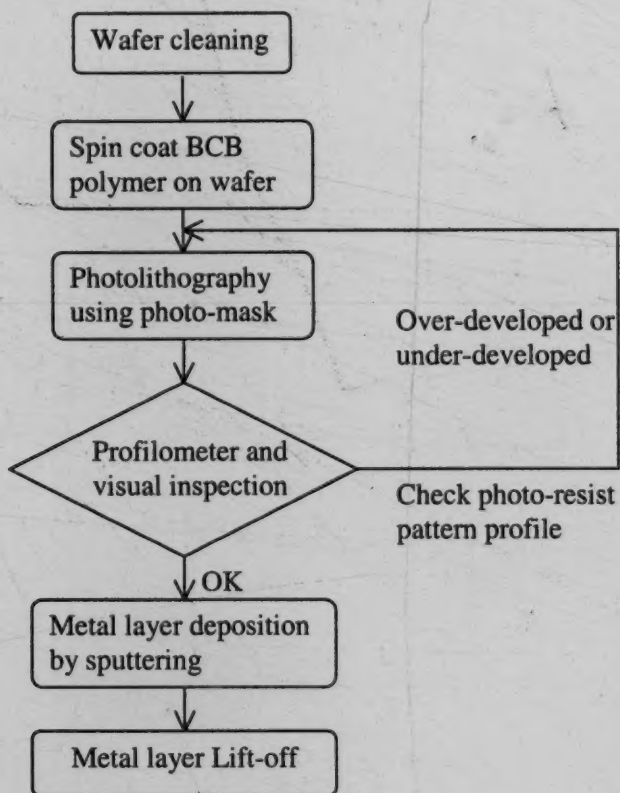


Fig. 4.4: Fabrication process flow chart

4.5. Experiment setup

A HP8510B Vector Network Analyzer (VNA) was connected to GSG-150-DP probes fixed on probe stations that are capable of 3D movement via its two ports.

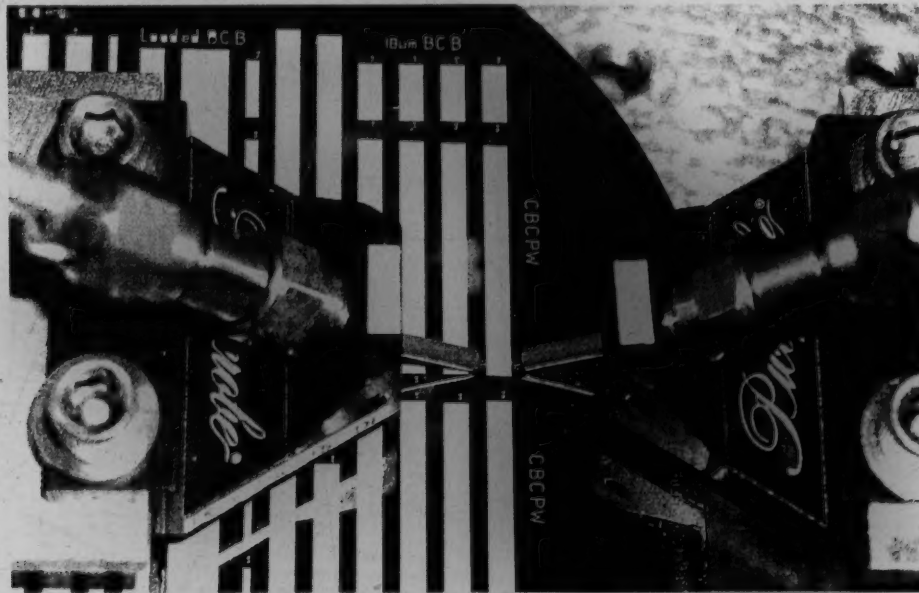


Fig. 4.5: Close-up view of probe on wafer measurement of CBCPW quarter-wavelength resonator

The figure shows the experiment setup and measurement of a CBCPW quarter-wavelength resonator fabricated on BCB layer on top of a gold conducting layer of 500 nanometer thick on a silicon wafer. The three tip-probes (Ground-Signal-Ground) were brought into contact with the two-port CBCPW structure from above the device. One probe was used on each end of the resonator structure in order to launch and terminate the applied signal.

5. Results analysis and discussions

5.1. CPW line on quartz

All CPW lines are 10 mm long and built on 580 μm thick quartz wafer with dimensions shown in Table 4.2.1.2 (a). The measured reflection and transmission losses in frequency domain are recorded in Fig. 5.1.1 to Fig. 5.1.3.

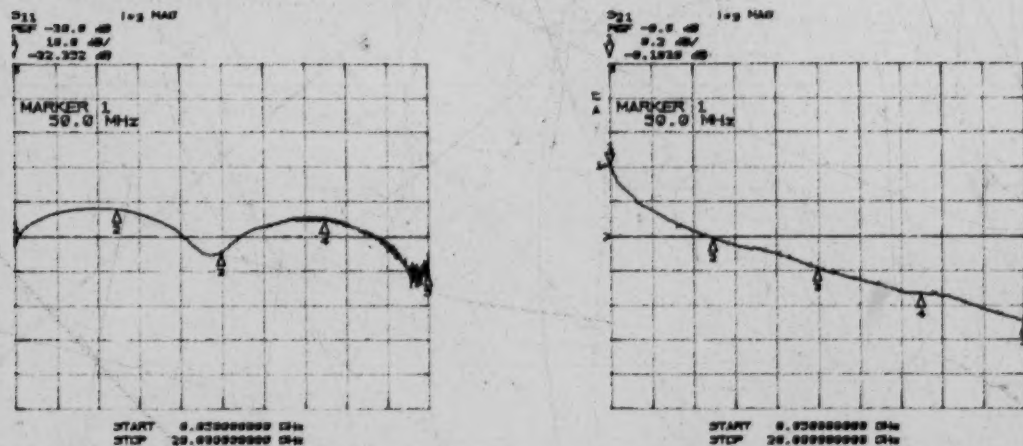


Fig. 5.1.1: Measured S_{11} and S_{21} in CPW line ($g=16 \mu\text{m}$)



Fig. 5.1.2: Measured S_{11} and S_{21} in CPW line ($g=12 \mu\text{m}$)

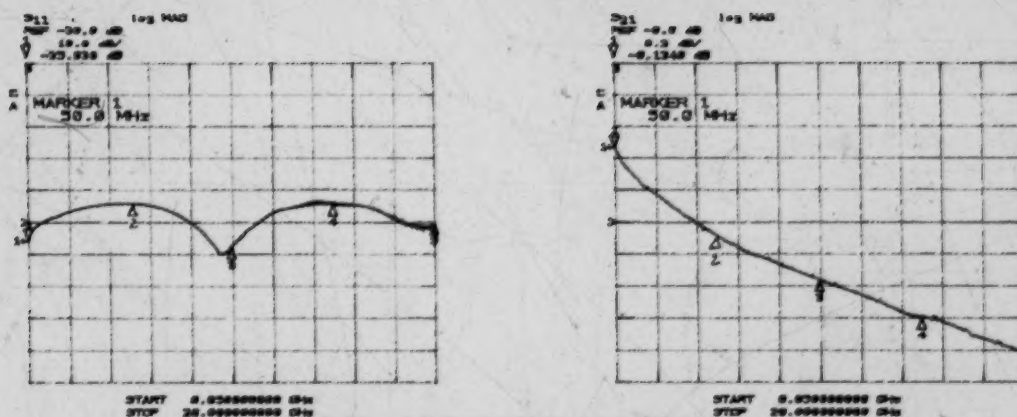


Fig. 5.1.3: Measured S_{11} and S_{21} in CPW line ($g=10 \mu\text{m}$)

These plots show that CPW lines with gap width $g=12 \mu\text{m}$ is best matched to 50Ω impedance, and the reflection loss caused by mismatch can be ignored (S_{11} is less than -25.5 dB across 20 GHz frequency range). The measured S_{21} curve closely matches the simulated conductor loss curve as shown in Fig. 5.1.4 for CPW on quartz structure with $g=12 \mu\text{m}$.

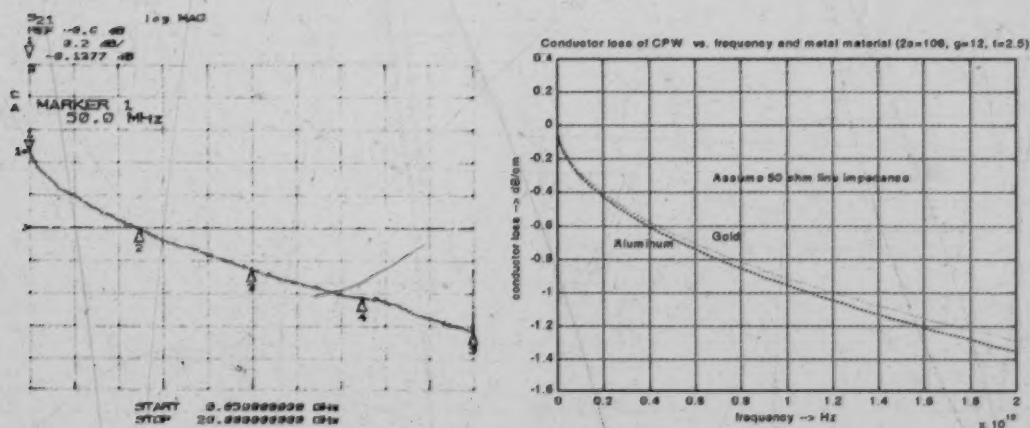


Fig. 5.1.4: Measured and simulated conductor loss in CPW line

The measured transmission losses in CPW lines with different gap widths and simulated conductor loss at five frequencies are recorded in the following table, the loss in CPW with gap width of 12 μm is most closely matched to the simulation due to the smallest impedance mismatch.

		50 MHz	4.9876 GHz	10.0000 GHz	15.0125 GHz	19.9501 GHz
Measured loss in CPW with $g=16 \pm 0.5 \text{ }\mu\text{m}$ (dB)		-0.1939	-0.6154	-0.7848	-0.9363	-1.0988
Measured loss in CPW with $g=12 \pm 0.5 \text{ }\mu\text{m}$ (dB)		-0.1377	-0.6145	-0.8557	-1.0351	-1.2393
Measured loss in CPW with $g=10 \pm 0.5 \text{ }\mu\text{m}$ (dB)		-0.1349	-0.6849	-0.9628	-1.1982	-1.4225
Simulated	Aluminum layer (dB)	-0.0673	-0.6733	-0.9522	-1.1466	-1.3467
	Gold layer (dB)	-0.0642	-0.6415	-0.9073	-1.0925	-1.2831

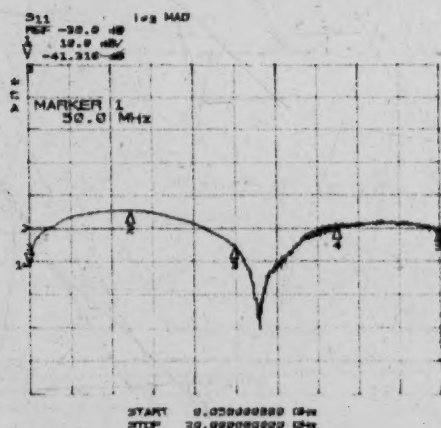
Table 5.1: Measured and simulated S_{21} in CPW line on quartz

The two simulated attenuation curves show the conductor loss in 50 Ω CPW with 2.5 μm aluminum and gold as the conductor respectively. The agreement between the measured and simulated results clearly show that conductor loss is the only loss mechanism in CPW on quartz structure in the investigated frequency range. The discrepancy between the measured loss and simulated conductor loss at 50 MHz is probably due to the unrealistic skin depth calculated for that frequency in the simulation. The skin depth is calculated to be thicker than the real conductor thickness and thus gave a smaller than real conductor loss at low frequency. This is why the increment inductance rule requires the conductor thickness at least two times of the skin depth. However, this discrepancy does not cause too much trouble since the losses are very small at low frequency. When the skin depth becomes smaller than the real conductor

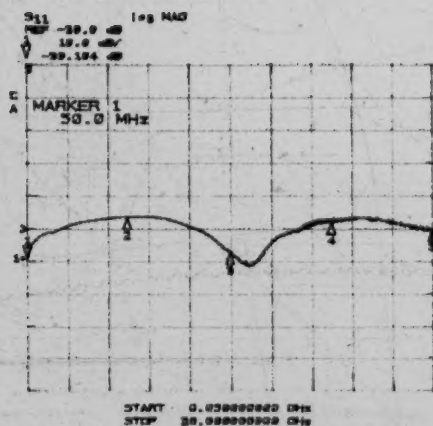
thickness at higher frequencies (2-3 GHz for most of good conductors), the simulated loss becomes closer to the real measurement and more accurate.

5.2. CPW line on BCB and quartz wafer

The S_{11} and S_{21} curves of 1 cm long CPW line on BCB and quartz wafer with gap width of 16 μm and 12 μm are shown in Fig. 5.2.



(a) Measured S_{11} and S_{21} in CPW line on BCB and quartz wafer ($g=16 \mu\text{m}$)



(b) Measured S_{11} and S_{21} in CPW line on BCB and quartz wafer ($g=12 \mu\text{m}$)

Fig. 5.2: Measured S_{11} and S_{21} in CPW line on BCB and quartz wafer

The measured transmission losses in CPW line on BCB on quartz wafer at 5 frequency points are recorded in the following table.

	50 MHz	4.9876 GHz	10.0000 GHz	15.0125 GHz	19.9501 GHz
$g=16\pm0.5 \mu\text{m}$ (dB)	-0.0765	-0.4593	-0.6026	-0.7520	-0.9155
$g=12\pm0.5 \mu\text{m}$ (dB)	-0.0897	-0.5701	-0.8023	-0.9953	-1.1936

Table 5.2: Measured S_{21} in 1cm CPW on BCB on quartz

The measurement obtained from CPW line on BCB structure show no significant difference from that of CPW line on quartz structure and the shape of S_{21} curves are largely the same, which means that BCB polymer layer does not add a noticeable dielectric loss to the total transmission loss and can be ignored compared to conductor loss. The slight decrease in transmission loss can be explained by the increase in the center conductor width and smaller effective dielectric constant in this structure compared with that of CPW on quartz structure as shown in Table 4.2.1.2 (a) and Table 4.2.1.2 (b).

5.3. CBCPW line with BCB polymer as dielectric layer

The transmission loss in CBCPW line is a key factor limiting the bandwidth of the EO polymer modulator [7], which is the focus of our investigation. The S_{11} and S_{21} curves of a 5 mm long and a 34.617 mm long 50 Ω CBCPW through line with center conductor width of 42 μm and gap width of 16 μm were measured respectively and the plots are shown in Fig. 5.3.1 and Fig. 5.3.2.

The reflection loss in the 5 mm long CBCPW line is below -21 dB, which means a good impedance match between the CBCPW line and the measurement system is achieved. The reflection loss is ignored, and the attenuation constant is given as:

$$\alpha = \frac{S_{21}}{8.686} \text{ (Neper / unit length)} \quad (5.3-1)$$

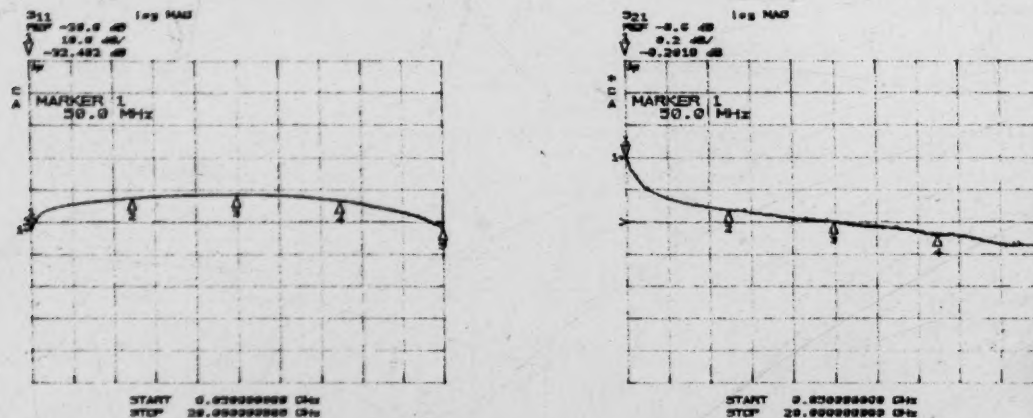


Fig. 5.3.1: Measured S_{11} and S_{21} in 5 mm long CBCPW line

The measured transmission loss and calculated attenuation constants at 5 frequencies in 5 mm long CBCPW line are recorded in Table 5.3.1.

Frequency	Measured S_{21} (dB)	α (Neper/5 mm)
50 MHz	-0.2010	0.02314
4.987 GHz	-0.528	0.06079
10 GHz	-0.5967	0.06870
14.987 GHz	-0.6787	0.07814
19.975 GHz	-0.7581	0.08728

Table 5.3.1: Measured S_{21} and calculated attenuation constant in 5 mm CBCPW

Using the attenuation constant values obtained in Table 5.3.1 and the following equation, we can calculate the transmission loss in 3.4617 cm long CBCPW through line as shown in Table 5.3.2.

$$P_{Transfer} = P_{Total} \cdot e^{-2\alpha l}, \quad (5.3-2)$$

where l is the length of the transmission line.

	50 MHz	4.987 GHz	10.000 GHz	14.987 GHz	19.975 GHz
Measured S_{21} in 5 mm CBCPW (dB)	-0.2010	-0.5280	-0.5967	-0.6787	-0.7581
Calculated S_{21} in 3.4617 cm CBCPW (dB)	-1.3915	-3.6556	-4.1313	-4.6990	-5.2486
Measured S_{21} in 3.4617 cm CBCPW (dB)	-1.0851	-3.0599	-3.7545	-4.4272	-4.9592

Table 5.3.2: Measured and calculated S_{21} in 3.4617 cm CBCPW

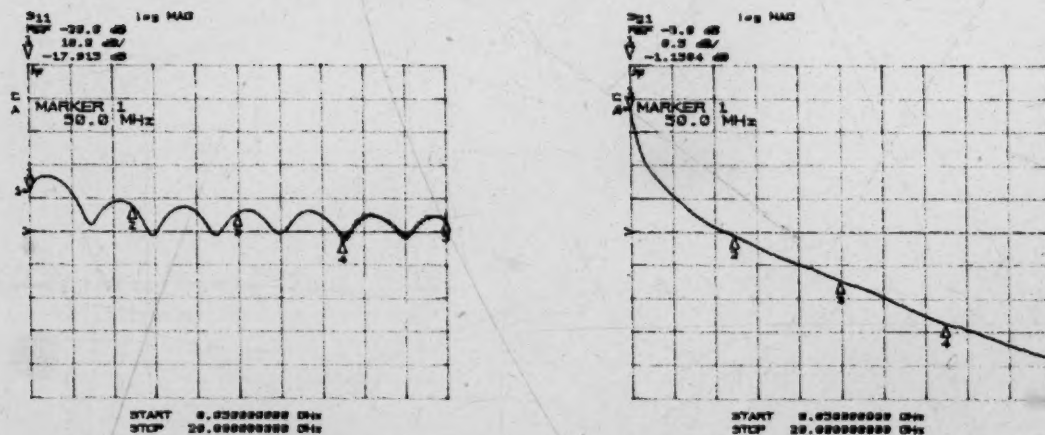


Fig. 5.3.2: Measured S_{11} and S_{21} in 34.617 mm long CBCPW line

The finite discrepancy between the calculated and measured transmission loss in 3.4617 cm CBCPW could be due to the finite mismatch between the measurement system and CBCPW line, small fabrication tolerances in the transmission line geometry (thus slightly different transmission character) and measurement errors.

The measured transmission loss in 3.4617 cm long CBCPW line ($g=16\text{ }\mu\text{m}$), simulated conductor loss in 3.4617 cm long CPW line on quartz ($g=12\text{ }\mu\text{m}$) and calculated transmission loss in the same CPW line at discrete frequency points by using the attenuation constant obtained in Table 5.1 are plotted in the following figure for a comparison.

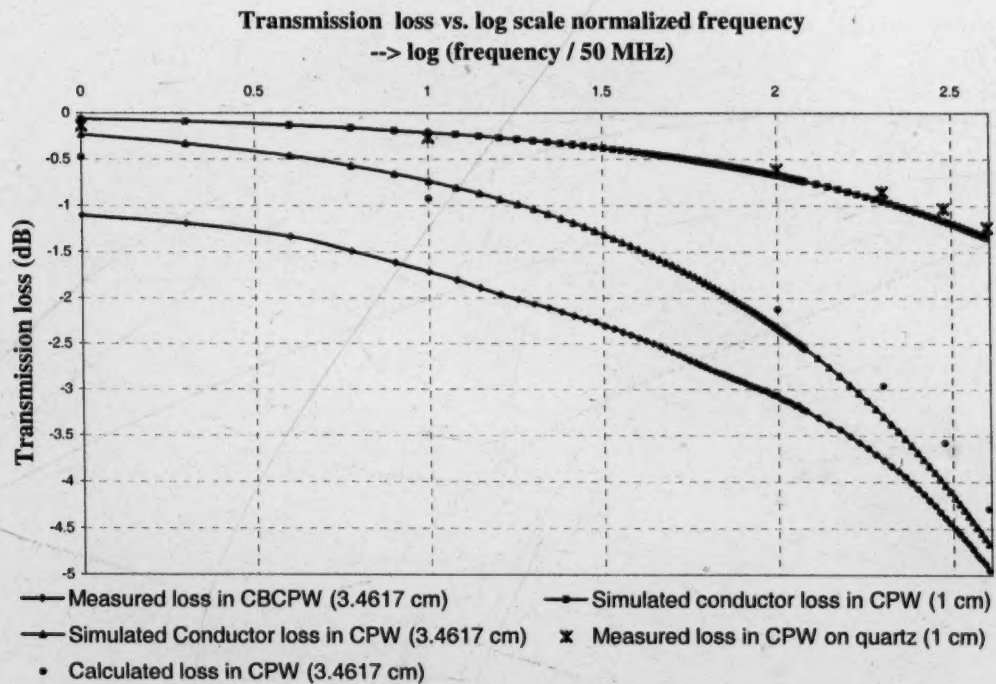


Fig. 5.3.3: Comparison of transmission loss in CPW and CBCPW

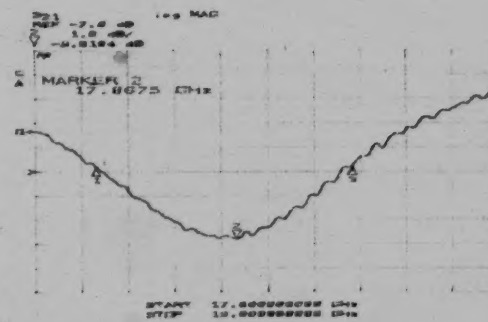
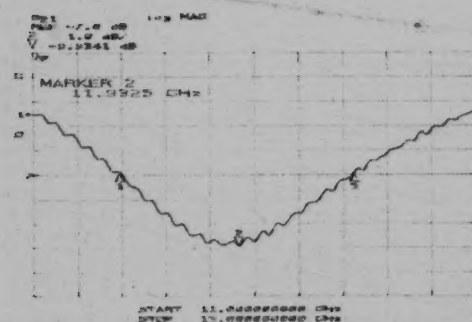
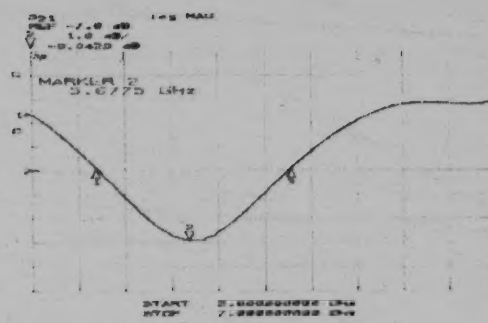
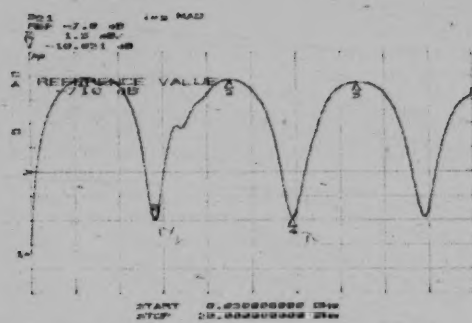
The top two loss curves are simulated conductor losses in 1 cm and 3.4617 cm long CPW lines respectively. The small cross marks along the top curve are measured conductor loss in 1 cm long CPW line on quartz. The extrapolated transmission loss in 3.4617 cm long CPW line at six frequencies were calculated by using the measured attenuation constants of 1 cm long CPW line on quartz and equation (5.3-2), and these points are marked as small dots along the simulated conductor loss curve in 3.4617 cm

CPW line. The measured transmission loss curve in CBCPW on the bottom of the plot closely follows the same trend as the extrapolated transmission loss in CPW on quartz of the same length. Since we have proved that the transmission loss in CPW on quartz structure is caused by conductor loss, the similarity between the extrapolated transmission loss in CPW line on quartz and the measured transmission loss in CBCPW of the same length indicating that conductor loss is still the dominant loss mechanism in CBCPW line. The bigger transmission loss in CBCPW compared to that of CPW line on quartz of the same length arises from its much narrower center conductor width compared to that of CPW line [48].

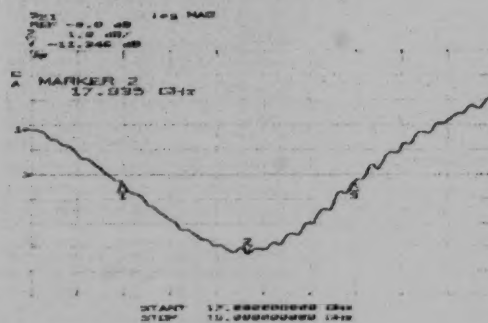
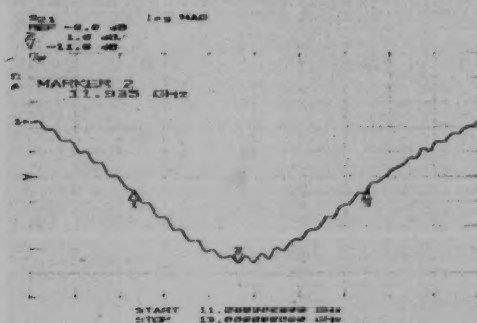
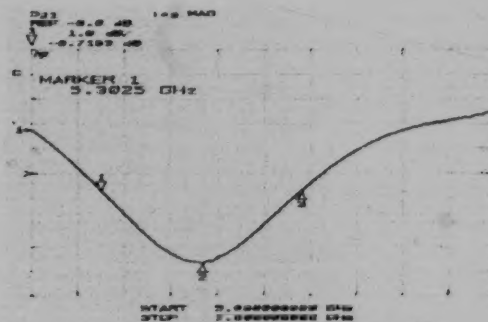
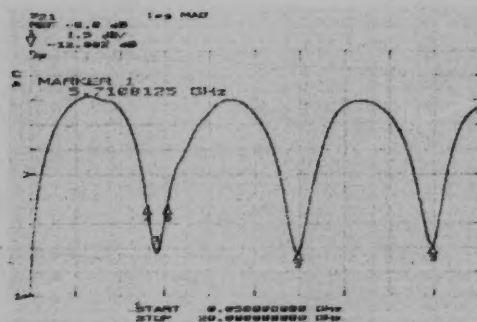
Since there is no accurate analytical model available to simulate conductor loss in CBCPW lines, quantitative analysis is very difficult. Never the less, the comparison between the above curves confirms the original speculation that large radiation or parallel-plate leakage losses in CBCPW structure are ruled out in the investigated frequency range since the transmission loss curve of CBCPW line closely follows the conductor loss curve of a CPW line in the whole frequency range investigated.

5.4. CBCPW resonator

The transmission characteristic of CBCPW resonators were measured, only the quarter-wavelength resonators gave good resonant curves. The resonance characteristics of two quarter-wavelength resonators with the same dimension (Center conductor width $2s=41\text{ }\mu\text{m}$, gap width $g=16\text{ }\mu\text{m}$, conductor thickness $t=2.5\text{ }\mu\text{m}$) are shown in Fig. 5.4.1.



(a) Resonator 1



(b) Resonator 2

Fig. 5.4.1: Resonator S_{21} measurements

The measured resonant frequencies and corresponding transmission loss figures and Q factors are recorded in Table 5.4.1. The calculated unloaded Q factors, effective dielectric constants and the attenuation constants are listed in Table 5.4.2 by using the equations in section 4.2.2.

	Resonator 1	Resonator 2
$f_{res\ 1}$ (GHz)	5.6775	5.74
$f_{res\ 2}$ (GHz)	11.9325	11.935
$f_{res\ 3}$ (GHz)	17.8675	17.935
$S_{21\ res1}$	-9.9429 dB (0.3183)	-11.708 dB (0.2598)
$S_{21\ res2}$	-9.9341 dB (0.3186)	-11.6 dB (0.2630)
$S_{21\ res3}$	-9.8184 dB (0.3229)	-11.346 dB (0.2708)
$Q_{res\ 1} (Q_L)$	6.759	6.6358
$Q_{res\ 2} (Q_L)$	11.257	11.5036
$Q_{res\ 3} (Q_L)$	16.169	17.935

Table 5.4.1: Measured S_{21} and loaded Q factor in two quarter-wavelength resonators with the same dimension

		Unloaded Q (Q_0)	Effective Dielectric Constant (ϵ_{eff})	Attenuation Constant (α_0)
Resonator 1	$f_{res\ 1}$ (5.6775 GHz)	7.569	2.330	1.04 dB/cm
	$f_{res\ 2}$ (11.9325 GHz)	12.609	2.110	1.25 dB/cm
	$f_{res\ 3}$ (17.8675 GHz)	18.175	2.117	1.30 dB/cm
Resonator 2	$f_{res\ 1}$ (5.74 GHz)	7.135	2.280	1.11 dB/cm
	$f_{res\ 2}$ (11.935 GHz)	12.393	2.110	1.27 dB/cm
	$f_{res\ 3}$ (17.935 GHz)	19.415	2.101	1.22 dB/cm

Table 5.4.2: Calculated unloaded Q factor, effective dielectric constant and attenuation constant at resonant frequencies in two quarter-wavelength resonators with the same dimension

The effective dielectric constant (ϵ_{eff}) of CBCPW shows less than 10% dispersion across the frequency range of 5.5 GHz to 18 GHz, and they are closely matched to the calculated static ϵ_{eff} of 2.086 by the quasi-static CBCPW model. There is no dispersion observed in the frequency range of 11.93 GHz to 17.9 GHz. Since the dispersion characteristic of CBCPW is mainly determined by that of BCB polymer, the results indicate that BCB polymer is a low dispersion dielectric suitable for wideband applications. The attenuation constant α_0 as a function of frequency obtained by using the matched-through-line method measurement of 5 mm CBCPW line and the resonance method are plotted in Fig. 5.4.2.

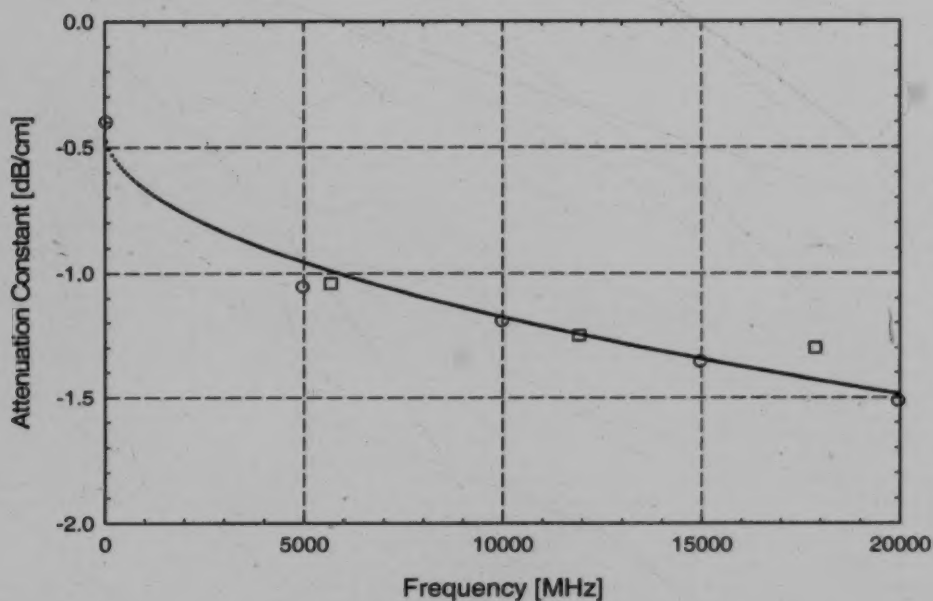


Fig. 5.4.2: Attenuation constant as a function of frequency obtained by the matched-through-line method and the resonance method

In Fig. 5.4.2, the circles and the squares are attenuation constants obtained by the matched-through-line method and the resonance method respectively. Both sets of data

generally follow the square-root-frequency-dependant conductor loss curve (See page 33) shown in the same plot which further proves that conductor loss is the dominant loss mechanism in CBCPW in the frequency range investigated. The attenuation constants obtained at the measured frequencies are summarized in Table 5.3.1.

Frequency	Through line measurement	Resonator 1 measurement
50 MHz	-0.402 dB/cm	
4.987 GHz	-1.056 dB/cm	
5.6775 GHz		-1.04 dB/cm
10 GHz	-1.1934 dB/cm	
11.9325 GHz		-1.25 dB/cm
14.987 GHz	-1.3574 dB/cm	
17.8675 GHz		-1.3 dB/cm
19.975 GHz	-1.5162 dB/cm	

Table 5.4.3: Comparison of attenuation constant α_0 obtained by the matched-through-line method and the resonance method

6. Conclusion and future works

6.1. BCB polymer as a low loss low dispersion dielectric material

The transmission loss and dispersion behavior of the BCB polymer were investigated by matched-through-line and quarter-wavelength resonance methods over a wide frequency range from 50 MHz to 20 GHz. The comparison between the transmission loss measurements in CPW on quartz and CPW on BCB configurations shows that BCB is a good dielectric material for low loss, wideband applications. Because the BCB dielectric layer does not introduce noticeable transmission loss to the CPW on BCB structure compared with the transmission loss in CPW on quartz structure of the same length. In the resonance method, it was also proved that dispersion in this polymer is very small across the wide frequency range, and there is virtually no dispersion effect observed in the frequency range of 11.9 GHz – 17.8 GHz in our experiment.

6.2. Radiation / Parallel-plate leaky loss in CBCPW line

A major concern over the employment of CBCPW configuration as the electrode structure for the modulator was related to the radiation / leakage loss imposed by the parallel-plate configuration in CBCPW. The measurements and simulations show that the parallel-plate leaky mode in CBCPW does not give rise to a noticeable transmission loss, and conductor loss due to the finite conductivity and skin depth effect is the major source of loss in CBCPW in our experiment. The dielectric loaded CBCPW structure is not needed to suppress the leakage problem stated in section 3.4.

6.3. Overall transmission loss of CBCPW line

We have demonstrated that the biggest transmission loss is 4.9 dB in a CBCPW through line of 3.6417 cm long at 20 GHz, which is much better than the worst-case requirement of 6.43 dB transmission loss to achieve half power modulation in the proposed modulator [7]. The major loss mechanism is conductor loss in the investigated frequency range. The reduction of the conductor loss could be achieved by employing methods such as using conductors with higher conductivities as the electrode material, moderately increasing the electrode thickness, increasing the center conductor width and reducing the conductor surface roughness.

In this experiment, Conductor-Backed coplanar Waveguide (CBCPW) is proved to be a viable electrode structure for the proposed broadband EO polymer modulator. Since CBCPW structure can be built on any planar dielectric surfaces without worrying about the electrical properties of the dielectric substrate, it is compatible with many dielectric systems. The transmission efficiency in CBCPW could be further improved by methods suggested above. CBCPW can be used as an economic circuit platform for high-speed electro-optic modulator applications.

6.4. Future work

The feasibility of deployment of CBCPW line as the electrode structure in polymer EO modulator on silicon substrate has been proved in this experiment in terms of its wideband transmission loss and dispersion. This has paved the way for continued research and development of silicon platform devices. The future work will be focused on the integration of active polymer optical waveguide with the CBCPW structure. The

next challenge for this proposed wideband EO polymer modulator is to find active polymer material as the optical waveguide that possesses properties such as high optical transparency, high EO coefficient and long term thermal stability, just to name a few, and fabrication related issues. The requirement of velocity matching between the optical and electrical signals in order to achieve a wide modulation bandwidth is another challenge, which might be achieved by adjusting the thickness of the polymer cladding layer for the optical waveguide.

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Appendix

A1. Transmission line fabrication process

- **Wafer cleaning**

The transmission line fabrication process begins with wafer cleaning, in which the wafers are chemically cleaned to remove particulate matter on the surface as well as any traces of organic, ionic, and metallic impurities. The wafers are put into a strong acid called Paranha, which is composed of three parts of H_2SO_4 and one part of H_2O_2 , for 20 minutes. Then the wafers are rinsed by de-ionized water and put into a convection oven baking at $110^\circ C$ for half an hour to dehydrate the wafers. Now, the smooth and highly polished wafer surface is ready for further process.

- **BCB polymer spin coating and thermal curing**

BCB-3022-63 manufactured by DowTM Chemical Company, was used as dielectric material layer in the transmission lines. It is a low viscosity clear polymer material, spin coating technique is commonly used as the deposition method.

An adhesion promoter was first spin coated on the wafer to improve the adhesion between the polymer layer and the wafer. Then, the wafer was put onto a spinner chuck and vacuum sucked on the chuck surface. First, a small amount of polymer was poured on the center of the wafer while the spinner was spinning at a low speed of 150 RPM, the amount of polymer used depends on the polymer thickness and the size of the wafer. For the polymer layer on my wafer, it was only about $18\ \mu m$ thick and 5 ml to 10 ml BCB was enough to cover the whole quartz wafer. Then the spinner was accelerated to an intermediate speed of 1500 RPM to spread out the polymer to cover the whole wafer

surface for about 20-30 seconds. Since the polymer tends to form a thicker bead at the out edge of the wafer, and is not desirable for good contact between the wafer and optic mask when photolithography is performed, a BCB striper to remove the edge bead was applied to the outer edge of the wafer at a spinning speed of 400-500 RPM for 10 seconds. Then to reach a final polymer layer thickness of $18\mu\text{m}$, the spinner had to be kept running at a speed of about 1600 RPM for 20 seconds. Then a hot plate was used to thermally cure the spin coated BCB layer, which was needed to evaporate the solvent in the polymer and cure it into a solid layer. This process needed a baking time of one hour at a temperature of 220°C .

Here, the advantage of hotplate baking over the conventional oven baking has to be mentioned. Since in hotplate baking process, the heating is from the wafer bottom to the top, solvent in polymer has a good opportunity to escape from the bottom of the layer through the top surface, and it won't suffer the skin effect in an convection oven environment, where the heating is applied from the outside in and solvent can be trapped inside the polymer layer by the formation of a 'hard skin' on polymer surface. In some series situation, the trapped solvent could cause peeling or bubbling in polymer layer. So hotplate baking is always a better choice.

- **Photolithography**

Photolithography is a critical process in any microstructure fabrications. It is the process of transferring geometric shapes on a mask to the surface of a wafer. In this process, a wafer coated by photoresist layer is brought into close contact with a top mask plate, and is exposed under a UV light source. The patterns on the mask then be

recorded onto the photoresist layer on the wafer, and the recorded pattern could be revealed by the process of developing as in a photo studio with photo resist developer.

The steps involved in photolithographic process are photoresist application; soft baking; mask alignment; exposure, development, and hard-baking.

- **Positive and Negative Photoresist**

There are two types of photoresist: positive and negative photoresist. For positive resists, the resist is exposed with UV light wherever the underlying material is to be removed. In these resists, exposure to the UV light changes the chemical structure of the resist so that it becomes more soluble in the developer. The exposed resist is then washed away by the developer solution, leaving windows of the bare underlying material. In other words, 'whatever shows, goes.' The mask, therefore, contains an exact copy of the pattern, which is to remain on the wafer. Negative resists behave in just the opposite manner. Exposure to the UV light causes the negative resist to become polymerized, and more difficult to dissolve. Therefore, the negative resist remains on the surface wherever it is exposed, and the developer solution removes only the unexposed portions of the resist. Masks used for negative photoresists, therefore, contain the inverse (or photographic "negative") of the pattern to be transferred. Negative resists were popular in the early history of integrated circuit processing, but positive resist gradually became more widely used since they offer better process controllability for small geometry features. Positive resists are now the dominant type of resist used in VLSI fabrication processes.

In our process, we first chose negative photoresist because the first mask was designed for negative photoresist, it was a clear field optic mask. The developing result was not promising, and the feature size had a large tolerance from the designed dimension. So the second mask was designed for positive photoresist, and it is a dark field mask.

- **Photoresist application**

Spin coating technique was used for photoresist application. The operation is similar to that of polymer coating process, except for the usage of different adhesion promoter for photoresist and different spinning speed depending on the desired thickness of the photoresist. The photoresist application was done in a clean room with a UV light controlled environment.

The thickness of the photoresist layer depends on the thickness of the metal layer that will be deposited. Theoretically, the thickness of the photoresist layer has to be greater than that of the metal layer to perform the subsequent lift-off operation. Otherwise, the steps formed between developed and undeveloped photoresist areas could be connected by the subsequently deposited metal layer, which could result in a lift-off failure. By estimation, we needed a greater than 3 μm photoresist layer to ensure the success of lift-off of a 3 μm metal layer. The photoresist I used in my process is STR™-1045 photoresist by SHIPLEY, which is a positive G-line (435 nm) and broadband photoresist engineered for thick resist applications. It is capable of forming thick photoresist layer ranging from 4 μm to 25 μm with a single coat process. Under optimum situation (exposure and development time as well as developer as

recommended in the data sheet) it can give a result of 1 μm lines/spaces with resist thickness of 5 μm .

- **Soft-baking**

Soft-baking is the step during which almost all of the solvents are removed from the photoresist coating. Soft-baking plays a very critical role in photo-imaging. The photoresist coatings become photosensitive, or imageable, only after soft-baking. Over soft-bake will degrade the photosensitivity of resists by either reducing the developer solubility or actually destroying a portion of the sensitizer. Under soft-bake will prevent light from reaching the sensitizer. Positive resists are incompletely exposed if considerable solvent remains in the coating. This under soft-baked positive resist is then readily attacked by the developer in both exposed and unexposed areas, causing less etching resistance. Although hotplate baking is a better choice, the hotplate in Microfab cannot be used to bake the photoresist. The size of the quartz wafer is bigger than the round hotplate available in the UV controlled clean room in MicroFab and the conventional oven had to be used. It required a gradual increase in the oven temperature and the wafers had to be baked for 30 minutes after the oven temperature reached 90-100° C degree.

- **Mask Alignment and Exposure**

One of the most important steps in the photolithography process is mask alignment. A mask or "photomask" is a square glass plate with a patterned emulsion of metal film on one side. The mask is aligned with the wafer, so that the pattern can be transferred onto the wafer surface. Each mask after the first one must be aligned to the

previous pattern if multi mask are to be used. Once the mask has been accurately aligned with the pattern on the wafer's surface, the photoresist is exposed through the pattern on the mask with a high intensity ultraviolet light. There are three primary exposure methods: contact, proximity, and projection. In Microfab, contact exposure is used.

- **Contact Printing**

In contact printing, the resist-coated wafer is brought into physical contact with the glass photo mask. The wafer is held on a vacuum chuck, and the whole assembly rises until the wafer and mask contact each other. The photoresist is exposed with UV light while the wafer is in contact position with the mask. Because of the contact between the resist and mask, very high resolution is possible in contact printing. The resolution of the final features also depends on the accuracy of the aligner, the exposure and developing time used. The problem with contact printing is that debris, trapped between the resist and the mask, can damage the mask and cause defects in the pattern.

- **Photoresist Development**

One of the last steps in the photolithographic process is photoresist development. In this process, after the photoresist is exposed, the wafer is put into photoresist developer, for the positive photoresist, the exposed photoresist will be washed away by a gentle agitation of the developer, while for the negative photoresist, the unexposed photoresist area will be washed away.

At low-exposure energies, the negative resist remains completely soluble in the developer solution. As the exposure is increased above a threshold energy E_t , more of the resist film remains after development. At exposures two or three times the threshold

energy, very little of the resist film is dissolved. For positive resists, the resist solubility in its developer is finite even at zero-exposure energy. The solubility gradually increases until, at some threshold, it becomes completely soluble. The quality of the photoresist development is affected by all the resist processing variables: initial resist thickness, soft-bake conditions, exposure time, developer chemistry, developing time, and others.

- **Hard-Baking**

Hard-baking is the final step in the photolithographic process. This step is necessary in order to harden the photoresist and improve adhesion of the photoresist to the wafer surface. It is suggested that this process be omitted if a lift-off will be performed later, which will make it easier to lift off the metal on top of the remaining photo resist.

- **Metal layer deposition**

There are several ways to deposit metal layer on a wafer, such as sputtering, electroplating and electric beam evaporation. Sputtering is chosen considering the controllability and availability of the facilities, because it is a well-developed process in the microfab facility and the sputtering system can support long sputtering time and big target is available to last long enough to yield $\sim 3 \mu\text{m}$ thick metal layer.

- **Metal layer lift-off**

This process is the final step for producing desired transmission line patterns, which works by putting the metal-coated wafer into ultrasonic acetone bath. The ultrasonic source produces numerous bubbles in acetone solvent, which explode omnidirectionally and release large amount of energy to loosen the bond between photoresist

layer and the wafer surface, acetone solvent slowly penetrate into and soften the photoresist, so the metal layer on top of photoresist would lose ground and be lift-off. The metal left is the desired transmission line pattern.